

1503/3

BULETINUL INSTITUTULUI POLITEHNIC DIN IAȘI

TOMUL XXX (XXXIV)

Fasc. 1—4

C: 040727

Secția III

**ELECTROTEHNICĂ
ENERGETICĂ
ELECTRONICĂ
AUTOMATIZĂRI**

1984

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Secția III

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BULETINUL INSTITUTULUI POLITEHNIC DIN IAȘI
BULLETIN OF THE POLYTECHNIC INSTITUTE OF JASSY
ИЗВЕСТИЯ ЯСКОГО ПОЛИТЕХНИЧЕСКОГО ИНСТИТУТА

Tom XXX (XXXIV), Fasc. 1-4

1984

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TEMPERATURE DEPENDENCE OF THE SEEBECK FACTOR AT Te—Bi—Sb—Se BASED SOLID SOLUTIONS IN BULK MATERIALS AND THIN FILMS

BY

ALEXANDRINA JEFLEA and SOFIA MELINTE

1. Introduction

The present work is dealing with the temperature dependence of the Seebeck factor at Bi_2Te_3 based thermoelectric materials, both as bulk materials and thin films. These materials are of interest in the production of thermoelectric generators, given their high factor of merit.

The measurements were made on samples with carrier concentrations between the intrinsic and degenerated regimes, at temperatures at which the materials present optimum working conditions.

Golds m i d [1] has shown that the parameter

$$(1) \quad F = \left(\frac{\mu_0}{K_L} \right) \left(\frac{m^*}{m} \right)^{3/2},$$

which, in a first approximation, does not depend on the Fermi level position, is a measure of the performances of the thermoelectric materials. In the relation (1), μ_0 is the non-degenerated mobility, K_L — the lattice constant in the thermal conductivity and m^* — the effective mass.

The maximum value of F is obtained for the compounds between $2(\text{Sb}_2\text{Te}_3) - (\text{Bi}_2\text{Te}_3)$ and $3(\text{Te}_3\text{Sb}_2) - (\text{Bi}_2\text{Te}_3)$ [2]. B i r k h o l t z studied the following materials: $\text{Bi}_2\text{Te}_3 - \text{Sb}_2\text{Te}_3$ and $\text{Bi}_2\text{Te}_3 - \text{Bi}_2\text{Se}_3$ and found that the reduction of the phononic component, which appears when increasing the Bi_2Se_3 amount, is accompanied by a decrease of electronic mobility [2], so that the plain effect on F is small. Taking these into account, the authors have chosen the following materials: 90% $\text{Bi}_2\text{Te}_3 - 10\% \text{Bi}_2\text{Se}_3$ and 25% $\text{Bi}_2\text{Te}_3 - 75\% \text{Sb}_2\text{Te}_3$, delivered in the corresponding concentrations by ICPIAF, Cluj-Napoca.

With the aim to use these materials as thin films in thermoelectric generators, the temperature dependence of the Seebeck coefficient for films at different thicknesses was studied in a first stage, as compared to the bulk material. Taking into consideration that at semiconductors

the Seebeck coefficient depends on the scattering of the charge carriers, its temperature dependence will be influenced by the thickness of the semiconductor layer, meaning that, besides the scattering mechanisms from the bulk materials, the scattering on the film surface will interfere.

2. Seebeck Coefficient

In the simple model of the spherical energetic zones, in the case of the dependence of the energy on the second power of the wave vector k and the isotropic relaxation time

(2)

$$\tau = \tau_0,$$

the relation giving the thermoelectric coefficient α within the classical statistics, it is for the free electronic gas in nondegenerated semiconductors, is [3]:

$$(3) \quad |\alpha| = \frac{k_0}{e} \left[r + \frac{5}{2} + \ln \frac{2(2\pi m^* k_0 T)}{h^3 n} \right],$$

where k_0 is the Boltzmann constant, h — the Planck constant, n — the carrier concentration; m^* — the carrier effective mass, r characterizes the energy dependence of the relaxation time and it is specific for each type of scattering.

Making a theoretical computation for Bi_2Te_3 by means of the relation (3), a value of $129 \mu\text{V}/\text{grd}$ was found for α , which is smaller than the experimental value; this disagreement was explained in [4] by the omission of the temperature variation of m^* after the law

$$m^* \sim T^s.$$

From the considerations concerning this dependence, Goltzman stated that the agreement between the values of s (-1.8 and 2) obtained from the measurements of the Seebeck coefficient and the conductivity confirms that the hypothesis about the prevailing part played by the acoustic scattering mechanism in the temperature and concentration range where the classical statistics is still valid.

According to these results, the value of r from (3) is $-1/2$ in the bulk materials, corresponding to the scattering on the acoustic phonons.

In the case of very thin films, where the thickness of the films is comparable with the carriers mean free path, r takes different values, corresponding to the scattering on the layer faces and discontinuities. Therefore, for very thin films, one can get for α higher values than those for the bulk materials.

3. The Experimental Procedure

Giving the difficulty of obtaining thin films made of these materials, more methods of fabrication were tried; finally, the vacuum flash evaporation was chosen, as being able to transfer the material from the evapor-

rator on the substrate without changing its composition. The device used for the vacuum flash evaporation permits that the material to be evaporated should get in the superheated evaporator in a small amount which instantaneously evaporates without dissociation, the condensation taking place on an unheated substrate [5]. The film thickness was determined by means of an interferometric microscope.

In order to determine the temperature dependence of α at the bulk materials, a device was made consisting of two copper parts whose temperatures can be changed by means of a heating resistance. The supply voltage can be adjusted so that the temperature difference between the two heaters should be 10°C . The temperature was measured with a copper — constantan thermocouple. Between the two metallic parts, the thermoelectric pellet is fixed which will therefore have the average temperature $\bar{T} = (T_c + T_r)/2 + T_{\text{med}}$, and the thermoelectric voltage U_s , measured on the pellet, corresponds to a temperature difference $\Delta T = T_c - T_r$. The Seebeck coefficient will be $\alpha = U_s/\Delta T$.

In order to measure the Seebeck coefficient at thin films, a device was made based on the same principle as the above one, but with some specific accomodations for thin films.

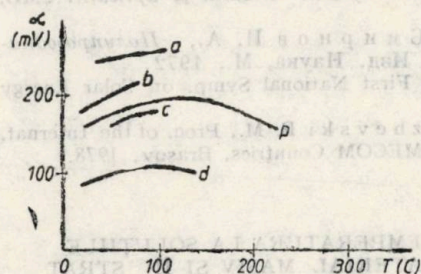


Fig. 1. — The temperature variation of Seebeck coefficient at a *p*-type material: *a* — 160 Å layer; *b* — 200 Å layer; *c* — 1 500 Å layer; *d* — 5 060 Å layer; *p* — the bulk material.

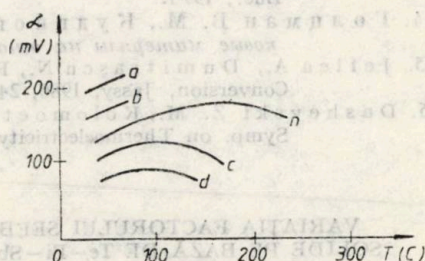


Fig. 2. — α vs. temperature for a *n*-type material: *a* — 150 Å layer; *b* — 190 Å layer; *c* — 2 250 Å layer; *d* — 4 000 Å layer; *n* — the bulk material.

The measurements were performed on pellets from different charges of thermoelectric materials and for films with the thicknesses ranging between 200 Å and 5 000 Å, deposited on cover glasses and on polyethylene terephthalate. The results of the measurements are presented in Figs. 1 and 2 for the materials of the *p* type (25% Bi_2Te_3 — 75% Sb_2Te_3) and *n*-type — (90% Bi_2Te_3 — 10% Bi_2Se_3) respectively.

4. Conclusions

As we can see in Figs. 1 and 2, the values of α for the bulk materials present a slight increase with increasing temperature due to the variation of the carrier concentration n , and a maximum between 90°C and 160°C

for the n -type material and a more evident maximum at $80^\circ\text{--}130^\circ\text{C}$ for a p -type material.

The p - and n -type thin layers present an increase of α with the decreasing film thickness. In the work [6], Kolomoetz indicated a 30% increase of α as compared to the bulk material.

The measurements for the very thin layers proved to be quite difficult, the layer becoming deteriorated at higher temperatures.

The most important conclusion of this work consists in the possibility of getting thin layers with improved thermoelectric properties, as compared to those of the bulk material, which will allow the successful utilization of thin layers in the thermoelectric devices.

Received, July 20, 1982

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VARIAȚIA FACTORULUI SEEBECK CU TEMPERATURA LA SOLUȚIILE SOLIDE PE BAZĂ DE Te-Bi-Sb-Se ÎN MATERIAL MASIV ȘI ÎN STRAT SUBȚIRE

(Rezumat)

Se studiază variația factorului Seebeck cu temperatura la materiale termoelectrice pe bază de Bi_2Te_3 , atât sub formă de material masiv cât și în strat subțire. Aceste materiale prezintă interes datorită factorului de merit ridicat, la utilizarea lor în generatoare termoelectrice.

4. Conclusions

QUELQUES OBSERVATIONS CONCERNANT DEUX THEOREMES DU CHAMP ELECTROMAGNETIQUE, EN MILIEUX NONLINEAIRES INERTIELS

PAR

HUGO ROSMAN

1. Introduction

Sont bien connues les difficultés heurtées lors de l'intégration du système d'équations du champ électromagnétique dans le cas des milieux nonlinéaires. Une possibilité de dépasser, dans une certaine mesure, ces difficultés est offerte par l'hypothèse du caractère „inertiel“ du milieu nonlinéaire, que l'on peut accepter souvent, en divers cas pratiques.

On admet que le milieu (nonlinéaire) est homogène, isotrope, sans polarisations électrique et magnétique permanentes et, toutefois, sans propriétés héréditaires de sorte qu'on peut écrire les lois de matériel du champ électromagnétique sous la forme :

$$(1) \quad \mathbf{D} = \varepsilon(E)\mathbf{E}, \quad \mathbf{J} = \sigma(E)\mathbf{E}, \quad \mathbf{B} = \mu(H)\mathbf{H}.$$

Si les „excitations“ $\mathbf{E} = \mathbf{E}(t)$, $\mathbf{H} = \mathbf{H}(t)$ sont des fonctions sinusoïdales de temps, les „réponses“ $\mathbf{D}(t)$, $\mathbf{J}(t)$, $\mathbf{B}(t)$ sont des fonctions périodiques de temps, mais nonsinusoïdales. On va dénommer comme „inertiel“ le milieu dont les caractéristiques en valeurs maxima (amplitudes), $D_m(E_m)$, $J_m(E_m)$, $B_m(H_m)$ sont nonlinéaires tandis que celles en valeurs instantanées $D(E)$, $J(E)$, $B(H)$ sont linéaires. Si l'on admet qu'on peut approximer les caractéristiques en amplitudes par des polynômes incomplets du troisième degré :

$$(2) \quad D_m(E_m) = \varepsilon_1 E_m - \varepsilon_3 E_m^3, \quad J_m(E_m) = \sigma_1 E_m - \sigma_3 E_m^3, \quad B_m(H_m) = \mu_1 H_m - \mu_3 H_m^3,$$

on obtient, pour les constantes de matériel (nonlinéaires) du milieu, les expressions [1] :

$$(3) \quad \varepsilon(E_m) = \varepsilon_1 - \frac{3}{4} \varepsilon_3 E_m^2, \quad \sigma(E_m) = \sigma_1 - \frac{3}{4} \sigma_3 E_m^2, \quad \mu(H_m) = \mu_1 - \frac{3}{4} \mu_3 H_m^2,$$

où E et H représentent les valeurs efficaces de l'intensité du champ électrique et, respectivement, magnétique.

Compte tenu que les caractéristiques $D_m(E_m)$, $J_m(E_m)$, $B_m(H_m)$ doivent être strictement croissantes, il résulte que :

$$(4) \quad \frac{dD_m}{dE_m} \geq 0, \quad \frac{dJ_m}{dE_m} \geq 0, \quad \frac{dB_m}{dH_m} \geq 0,$$

de sorte que les approximations (2) sont valables seulement dans les domaines :

$$(5) \quad E_m \leq \sqrt{\frac{\varepsilon_1}{3\varepsilon_3}} \quad \text{ou} \quad E_m \leq \sqrt{\frac{\sigma_1}{3\sigma_3}}, \quad H_m \leq \sqrt{\frac{\mu_1}{3\mu_3}}.$$

Par conséquent, l'approximation „inertielle“ du milieu nonlinéaire est valable seulement pour le cas de champs relativement faibles. Toutefois les constantes de matériel peuvent prendre des valeurs seulement dans les domaines :

$$(6) \quad \varepsilon \in \left[\frac{3}{4} \varepsilon_1, \varepsilon_1 \right], \quad \sigma \in \left[\frac{3}{4} \sigma_1, \sigma_1 \right], \quad \mu \in \left[\frac{3}{4} \mu_1, \mu_1 \right].$$

L'hypothèse du caractère „inertiel“ du milieu influence le contenu et donc la formulation de certains théorèmes concernant le champ électromagnétique, avec nombre de conséquences intéressantes qui seront relevées dans ce qui suit. Dans ce but on va se rapporter, d'une façon concrète, au théorème de la relaxation de la charge électrique libre et au théorème de la refraction des lignes de courant.

2. Théorème de la relaxation de la charge libre

On sait [2] que dans un corp de permittivité ε et de conductibilité σ , la densité de la charge électrique libre varie en fonction du temps selon la loi exponentielle :

$$(7) \quad \rho_v(t) = \rho_v(t_0) e^{-t/\tau_r},$$

où

$$(8) \quad \tau_r = \frac{\varepsilon}{\sigma}$$

représente la constante de relaxation. Pour les milieux nonlinéaires :

$$(9) \quad \tau_r = \tau_r(E) = \frac{\varepsilon(E)}{\sigma(E)}$$

et si le milieu peut être considéré comme „inertiel“, sur la base des relations (3) on obtient :

$$(10) \quad \tau_r(E_m) = \frac{\varepsilon_1 - \frac{3}{4} \varepsilon_3 E_m^2}{\sigma_1 - \frac{3}{4} \sigma_3 E_m^2}.$$

Or

$$(11) \quad \frac{d\tau_r}{dE_m} = \frac{3}{2} \cdot \frac{\varepsilon_1 \sigma_3 - \varepsilon_3 \sigma_1}{\left(\sigma_1 - \frac{3}{4} \sigma_3 E_m^2 \right)^2} E_m \geq 0.$$

On constate facilement que la fonction $\tau_r(E_m)$ est monotone, à savoir, strictement croissante si $\varepsilon_1/\varepsilon_3 > \sigma_1/\sigma_3$ et strictement décroissante dans le cas opposé. Dans la situation particulière où

$$(12) \quad \frac{\varepsilon_1}{\varepsilon_3} = \frac{\sigma_1}{\sigma_3}$$

la constante de relaxation devient

$$(13) \quad \tau_r = \frac{\epsilon_1}{\sigma_1} = \frac{\epsilon_3}{\sigma_3}$$

en étant indépendante de E_m .

Il est nécessaire à observer que, en outre, si $\epsilon_1/\epsilon_3 > \sigma_1/\sigma_3$, $E_m \leq \sqrt{\sigma_1/3\sigma_3}$ et si $\epsilon_1/\epsilon_3 < \sigma_1/\sigma_3$, $E_m \leq \sqrt{\epsilon_1/3\epsilon_3}$.

3. Le théorème de la refraction des lignes de courant

En régime stationnaire ou quasistationnaire, dans n'importe quel point d'une surface de discontinuité située dans un milieu conducteur, la relation [2]:

$$(14) \quad \sigma_1 E_{1n} = \sigma_2 E_{2n}$$

est toujours satisfaite. Si le milieu est faiblement conducteur et les permittivités des domaines séparés par la surface de discontinuité sont ϵ_1 , ϵ_2 alors, dans le même point est satisfaite et la relation:

$$(15) \quad \epsilon_2 E_{2n} - \epsilon_1 E_{1n} = \rho_s$$

où ρ_s est la densité superficielle de la charge électrique libre et E_{1n} , E_{2n} — les composantes normales de l'intensité du champ électrique. Dans le cas des milieux linéaires la condition nécessaire et suffisante pour que la surface de discontinuité soit nonchargée est que les constantes de relaxation des deux milieux soient égales [2] ($\epsilon_1/\sigma_1 = \epsilon_2/\sigma_2$). Si les milieux sont, tous les deux, nonlinéaires „inertiels“:

$$(16) \quad \begin{cases} \sigma_1 = \sigma_{11} - \frac{3}{4} \sigma_{13} E_{1m}^2, & \sigma_2 = \sigma_{21} - \frac{3}{4} \sigma_{23} E_{2m}^2, \\ \epsilon_1 = \epsilon_{11} - \frac{3}{4} \epsilon_{13} E_{1m}^2, & \epsilon_2 = \epsilon_{21} - \frac{3}{4} \epsilon_{23} E_{2m}^2, \end{cases}$$

de sorte que les relations (14) et (15) deviennent:

$$(17) \quad \begin{cases} \left(\sigma_{11} - \frac{3}{4} \sigma_{13} E_{1m}^2 \right) E_{1m} \cos \alpha_1 = \left(\sigma_{21} - \frac{3}{4} \sigma_{23} E_{2m}^2 \right) E_{2m} \cos \alpha_2, \\ \left(\epsilon_{21} - \frac{3}{4} \epsilon_{23} E_{2m}^2 \right) E_{2m} \cos \alpha_2 - \left(\epsilon_{11} - \frac{3}{4} \epsilon_{13} E_{1m}^2 \right) E_{1m} \cos \alpha_1 = \rho_s \end{cases}$$

et l'annulation de la densité superficielle de la charge électrique a lieu seulement pour certaines valeurs des amplitudes E_{1m} , E_{2m} des intensités du champ

¹⁾ On a admis que les oscillations de l'intensité du champ électrique dans les deux milieux sont symphasiques, hypothèse complètement acceptable.

électrique. La détermination de ces valeurs est assez difficile mais dans certains cas particuliers s'avère relativement simple. Ce sont les cas où les milieux ont seulement soit la conductibilité soit la permittivité nonlinéaire. Dans le deuxième cas les relations (17) deviennent :

$$(18) \quad \begin{cases} \sigma_1 E_{1m} \cos \alpha_1 = \sigma_2 E_{2m} \cos \alpha_2, \\ \left(\varepsilon_{21} - \frac{3}{4} \varepsilon_{23} E_{2m}^2 \right) E_{2m} \cos \alpha_2 - \left(\varepsilon_{11} - \frac{3}{4} \varepsilon_{13} E_{1m}^2 \right) E_{1m} \cos \alpha_1 = \rho_s. \end{cases}$$

Si l'on élimine de ces relations, par exemple E_{2m} , pour la densité superficielle ρ_s résulte l'expression :

$$(19) \quad \rho_s = \frac{E_{1m} \cos \alpha_1}{\sigma_2^2 \cos^2 \alpha_2} \left[\sigma_2^2 \cos^2 \alpha_2 (\varepsilon_{21} \sigma_1 - \varepsilon_{11} \sigma_2) - \frac{3}{4} (\varepsilon_{23} \sigma_1^3 \cos^2 \alpha_1 - \varepsilon_{13} \sigma_2^3 \cos^2 \alpha_2) E_{1m}^2 \right].$$

On peut déduire que ρ_s s'annule seulement pour certaines valeurs particulières des intensités du champ électrique dans les deux milieux à savoir :

$$(20) \quad E_{1m0} = \frac{2}{\sqrt{3}} \sigma_2 \sqrt{\frac{\varepsilon_{21} \sigma_1 - \varepsilon_{11} \sigma_2}{\varepsilon_{23} \sigma_1^3 \cos^2 \alpha_1 - \varepsilon_{13} \sigma_2^3 \cos^2 \alpha_2}} \cos \alpha_2$$

et

$$(21) \quad E_{2m0} = \frac{2}{\sqrt{3}} \sigma_1 \sqrt{\frac{\varepsilon_{21} \sigma_1 - \varepsilon_{11} \sigma_2}{\varepsilon_{23} \sigma_1^3 \cos^2 \alpha_1 - \varepsilon_{13} \sigma_2^3 \cos^2 \alpha_2}} \cos \alpha_1.$$

Évidemment, pour que la densité superficielle, ρ_s , s'annule il est nécessaire que

$$(22) \quad \text{sign} (\varepsilon_{21} \sigma_1 - \varepsilon_{11} \sigma_2) = \text{sign} (\varepsilon_{23} \sigma_1^3 \cos^2 \alpha_1 - \varepsilon_{13} \sigma_2^3 \cos^2 \alpha_2).$$

Donc, par opposition au cas des milieux linéaires où l'annulation de la densité superficielle de la charge sur une surface de discontinuité est assurée par une relation entre les constantes de matériel des deux milieux, en étant indépendante de l'intensité du champ électrique, dans les milieux faiblement conducteurs, à permittivité nonlinéaire „inertielle“, cette condition est réalisée seulement pour certaines valeurs, déterminées, de l'intensité du champ électrostatique dans les deux milieux. Les valeurs respectives sont fonctions non seulement des constantes de matériel des milieux mais aussi et de l'orientation des vecteurs de l'intensité du champ électrique par rapport à la surface de discontinuité.

Si seulement l'un des deux milieux est nonlinéaire (par exemple le milieu 1 et par conséquent $\sigma_{23}=0$, $\varepsilon_{23}=0$), alors

$$(23) \quad E_{1m0} = \frac{2}{\sqrt{3}} \sqrt{\frac{\varepsilon_{11} \sigma_2 - \varepsilon_{21} \sigma_1}{\varepsilon_{13} \varepsilon_2}}, E_{2m0} = \frac{2}{\sqrt{3}} \cdot \frac{\sigma_1 \cos \alpha_1}{\sigma_2 \cos \alpha_2} \sqrt{\frac{\varepsilon_{11} \sigma_2 - \varepsilon_{21} \sigma_1}{\varepsilon_{13} \sigma_2}}.$$

Il est intéressant à observer que tandis que la valeur de E_{1m0} ne dépend pas de l'orientation des vecteurs du champ électrique dans les deux milieux, la valeur de E_{2m0} en dépend. En même temps la condition (22) devient

$$(24) \quad \frac{\varepsilon_{11}}{\varepsilon_{21}} > \frac{\sigma_1}{\sigma_2}.$$

Selon ce qu'on connaît les constantes de relaxation de deux milieux linéaires homogènes, isotropes, sans polarisation électrique permanente, sont égales si la surface de séparation n'est pas chargée. Si les deux milieux sont nonlinéaires mais „inertiels“, l'égalité des constantes de relaxation se conserve dans les mêmes conditions. Dans une telle situation on peut déterminer les constantes de relaxation à l'aide des relations :

$$(25) \quad \tau_{r1} = \frac{\varepsilon_{11} - \frac{3}{4} \varepsilon_{13} E_{1m}^2}{\sigma_{11} - \frac{3}{4} \sigma_{13} E_{1m}^2}, \quad \tau_{r2} = \frac{\varepsilon_{21} - \frac{3}{4} \varepsilon_{23} E_{2m}^2}{\sigma_{21} - \frac{3}{4} \sigma_{23} E_{2m}^2}.$$

Si l'on prend en considération le cas le plus simple où les milieux ont seulement la permittivité nonlinéaire on obtient :

$$(26) \quad \tau_{r1} = \frac{\varepsilon_{11} - \frac{3}{4} \varepsilon_{13} E_{1m}^2}{\sigma_1}, \quad \tau_{r2} = \frac{\varepsilon_{21} - \frac{3}{4} \varepsilon_{23} E_{2m}^2}{\sigma_2}$$

et en substituant E_{1m} , E_{2m} avec les expressions (20), (21) on obtient :

$$(27) \quad \tau_{r1} = \tau_{r2} = \frac{\varepsilon_{13} \varepsilon_{21} \sigma_2^2 \cos^2 \alpha_2 - \varepsilon_{11} \varepsilon_{23} \sigma_1^2 \cos^2 \alpha_1}{\varepsilon_{13} \sigma_2^3 \cos^2 \alpha_2 - \varepsilon_{23} \sigma_1^3 \cos^2 \alpha_1}.$$

Par une voie analogue on peut étudier et le cas particulier où les deux milieux nonlinéaires ont la conductibilité nonlinéaire et la permittivité linéaire ; les conclusions sont tout à fait similaires.

4. Conclusions

Les théorèmes de la relaxation et de la refraction des lignes de courant ont, dans des milieux nonlinéaires „inertiels“, certaines particularités à savoir :

1. La constante de relaxation est une fonction monotone de l'amplitude de l'intensité du champ électrique. Seulement dans certains cas particuliers cette constante est indépendante de l'intensité du champ électrique.

2. On peut réaliser l'annulation de la densité superficielle de la charge sur la surface de discontinuité seulement pour certaines valeurs des amplitudes du champ électrique dans les deux milieux. Ces valeurs dépendent tant des constantes de matériel des milieux que de l'orientation des vecteurs des intensités des champs électriques par rapport à la surface de discontinuité.

Reçue le 29 novembre 1983

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CÎTEVA OBSERVAȚII REFERITOARE LA DOUĂ TEOREME ALE CÎMPULUI ELECTROMAGNETIC, ÎN MEDII NELINIARE INERȚIALE

(Rezumat)

În medii neliniare „inerțiale” teorema relaxației sarcinii electrice libere și teorema refracției liniilor de curent au anumite particularități, dintre care cele mai interesante sînt următoarele :

1. Constanta de relaxație este o funcție monotonă de amplitudinea intensității cîmpului electric. Numai în anumite cazuri particulare această constantă este independentă de intensitatea cîmpului electric exterior.
2. Anularea densității superficiale de sarcină de pe suprafața de discontinuitate este realizată numai pentru valori precizate ale intensității cîmpului electric în cele două medii. Aceste valori depind atît de constantele de material ale mediilor cît și de orientarea vectorilor intensităților cîmpului electric în raport cu suprafața de discontinuitate.

4. Conclusions

Les théorèmes de la relaxation et de la refraction des lignes de courant ont, dans des milieux nonlinéaires „inertiels”, certaines particularités :

1. La constante de relaxation est une fonction monotone de l'amplitude de l'intensité du champ électrique. Seulement dans certains cas particuliers cette constante est indépendante de l'intensité du champ électrique.

2. On peut réaliser l'annulation de la densité superficielle de la charge sur la surface de discontinuité seulement pour certaines valeurs des amplitudes du champ électrique dans les deux milieux. Ces valeurs dépendent tant des constantes de matériel des milieux que de l'orientation des vecteurs des intensités des champs électriques par rapport à la surface de discontinuité.

INTEGRAL TRANSFORMS ON ALGEBRAS (II)

BY

DANIEL CONDURACHE

1. An integral transformation on Sobrero algebra is proposed, taking into account the statements from [2].

2. Let be the 4th order algebra associative and commutative over the real number field with the multiplication table :

(1)

	ϵ_1	ϵ_2	ϵ_3	ϵ_4
ϵ_1	ϵ_1	ϵ_2	ϵ_3	ϵ_4
ϵ_2	ϵ_2	0	ϵ_4	0
ϵ_3	ϵ_3	ϵ_4	$-\epsilon_1$	$-\epsilon_2$
ϵ_4	ϵ_4	0	$-\epsilon_2$	0

We shall denote this algebra as $\mathcal{G}$ [5], [1].

With denotations $\epsilon_1=1$, $\epsilon_2=\epsilon$, $\epsilon_3=j$, $\epsilon_4=j\epsilon$, table (1) becomes :
 $1 \cdot \epsilon = \epsilon \cdot 1 = \epsilon$, $j \cdot 1 = 1 \cdot j = j$, $\epsilon \cdot j = j \cdot \epsilon = j\epsilon$, $\epsilon^2=0$, $j^2=-1$, $1^2=1$.

A certain element of algebra $\mathcal{G}$:

$$(2) \quad \hat{w} = a_1 + a_2\epsilon + a_3j + a_4j\epsilon; \quad a_i \in \mathbb{R}, \quad (i=\overline{1, 4}),$$

represents symbolically, with the derivation operator :

$$(3) \quad \hat{d} = \delta + \epsilon + j\omega; \quad \delta, \omega \in \mathbb{R},$$

signals like :

$$(4) \quad s(t) = \mathcal{A}(t) \sin [\omega t + \theta(t)],$$

where :

$$\mathcal{A}(t) = e^{\delta t} \sqrt{(a_2 + a_4 t)^2 + (a_1 + a_3 t)^2}, \quad \theta(t) = \arctan \frac{a_2 + a_4 t}{a_1 + a_3 t}.$$

Elements (2) characterized by :

(4₁)

$$a_1 a_4 = a_2 a_3$$

represent symbolically pure amplitude modulated signals and shall be denoted, as it follows, as „special-type elements“.

Signals like (4) shall be demonstrated to have the remarkable propriety of constituting bases in order to synthesise a large class of signals (i.e. any signal in $\mathbb{L}^1$).

3. The set of hypercomplex functions of two real variables shall be denoted as $\mathbb{F} = \{\hat{f} \mid \hat{f}: \mathbb{R} \times \mathbb{R} \rightarrow \mathcal{G}\}$ and

$$\mathbb{L}_1^1 = \left\{ f: \mathbb{R} \rightarrow \mathbb{R} \mid \int_{-\infty}^{\infty} t^k f(t) dt < \infty, (k=0,1) \right\}.$$

Let be function $\hat{\mathbb{F}}: \mathbb{L}_1^1 \rightarrow \mathbb{F}$ given by the integral transformation:

$$(5) \quad \hat{\mathbb{F}}_f = \int_{-\infty}^{\infty} f(t) e^{-(\delta + \varepsilon + j\omega)t} dt.$$

Theorem 1. *Integral transformation (5) exists for any function in $\mathbb{L}_1^1$, and is a monogenic one if considered as a hypercomplex function of a hypercomplex variable.*

Proof. Taking into account the definition of the exponential on a commutative algebra [5] and (1), the following separation is obtained:

$$(6) \quad \begin{aligned} e^{-(\delta + \varepsilon + j\omega)t} &= e^{-\delta t} e^{-j\omega t} e^{\varepsilon t} = e^{-\delta t} (\cos \omega t - j \sin \omega t) \sum_{k=0}^{\infty} \frac{(-1)^k \varepsilon^k t^k}{k!} = \\ &= e^{-\delta t} (\cos \omega t - j \sin \omega t) (1 - t\varepsilon) = \\ &= e^{-\delta t} (\cos \omega t - t \cos \omega t \cdot \varepsilon - \sin \omega t \cdot j + t \sin \omega t \cdot j\varepsilon). \end{aligned}$$

Applying the additivity of the integral and the above decomposition, the first part of the enunciation is obtained.

For its second part, the exterior product should be calculated [5]:

$$\hat{\mathbb{F}}_f \wedge d(\varepsilon + j\omega) = \int_{-\infty}^{\infty} f(t) [e^{-(\delta + \varepsilon + j\omega)t} \wedge d(\delta + \varepsilon + j\omega)] dt,$$

which is vanish identically in algebra $\mathcal{G}$.

Remarks. a) To specify $\hat{\mathbb{F}}_f$ for a certain function, in $\mathbb{L}_1^1$, means to specify for real functions of two scalar variables denoted as „components“ of $\hat{\mathbb{F}}_f$:

$$\hat{\mathbb{F}}_f = F_1(\delta, \omega) + F_2(\delta, \omega)\varepsilon + F_3(\delta, \omega)j + F_4(\delta, \omega)j\varepsilon.$$

The monogeneity supposes the existence of some differential relations among the components of $\hat{\mathbb{F}}_f$ [5].

b) Algebra $\mathcal{S}$ contains a subalgebra which is isomorphic with the complex field (i.e. the algebra generated by the family $\{\lambda + \mu\}_{\lambda, \mu \in \mathbb{R}}$).

We shall denote as $\mathcal{L}_f(s)$ the Laplace transform of function f on this subalgebra. Taking into account (5) and the known properties of Laplace transform, we obtain:

$$\begin{aligned}
 \hat{\mathbb{F}}_f &= \int_{-\infty}^{\infty} f(t) e^{-(\delta + \varepsilon + j\omega)t} dt = \int_{-\infty}^{\infty} f(t) e^{-(\delta + j\omega)t} e^{-\varepsilon t} dt = \\
 (7) \quad &= \int_{-\infty}^{\infty} f(t) e^{-(\delta + j\omega)t} (1 - \varepsilon t) dt = \int_{-\infty}^{\infty} f(t) e^{-(\delta + j\omega)t} dt - \varepsilon \int_{-\infty}^{\infty} t f(t) e^{-(\delta + j\omega)t} dt = \\
 &= \mathcal{L}_f(\delta + j\omega) + \varepsilon \left. \frac{d\mathcal{L}_f(s)}{ds} \right|_{s=\delta + j\omega}
 \end{aligned}$$

Using separation (7) and the known properties of Laplace transform, we may demonstrate:

Theorem 2. The integral transformation (5) is an $\mathbb{R}$ -linear one and has the properties:

$$\begin{aligned}
 (8) \quad \hat{\mathbb{F}}_{d^k f} &= (\delta + \varepsilon + j\omega)^k \hat{\mathbb{F}}_f, \quad k \in \mathbb{N}; \\
 \hat{\mathbb{F}}_{f * g} &= \hat{\mathbb{F}}_f \cdot \hat{\mathbb{F}}_g, \quad \hat{\mathbb{F}}_{\delta(t)} = 1,
 \end{aligned}$$

where $*$ denotes the convolution product and $\delta(t)$ is Dirac function.

Theorem 3. Transformation (5) is invertible and its inverse is:

$$(9) \quad \hat{\mathbb{F}}_f^{-1}(t) = \frac{f(t+0) + f(t-0)}{2} = \frac{1}{2\pi j} \int_{\mathcal{S}} \hat{\mathbb{F}}_f e^{\hat{a}t} d\hat{a}, \quad \hat{a} = \delta + \varepsilon + j\omega.$$

Proof. We successively obtain;

$$\begin{aligned}
 &\frac{1}{2\pi j} \int_{\mathcal{S}} \hat{\mathbb{F}}_f e^{\hat{a}t} d\hat{a} = \frac{1}{2\pi j} e^{\varepsilon t} \int_{\mathcal{S}} \hat{\mathbb{F}}_f e^{(\delta + j\omega)t} d(\delta + j\omega) = \\
 &= \frac{1 - \varepsilon t}{2\pi j} \int_{\mathcal{S}} \left[\mathcal{L}_f(\delta + j\omega) + \varepsilon \left. \frac{d\mathcal{L}_f(s)}{ds} \right|_{s=\delta + j\omega} \right] e^{(\delta + j\omega)t} d(\delta + j\omega) = \\
 &= \frac{1 - \varepsilon t}{2\pi j} \int_{\mathcal{S}} \mathcal{L}_f(\delta + j\omega) e^{(\delta + j\omega)t} d(\delta + j\omega) + \frac{\varepsilon(1 - \varepsilon t)}{2\pi j} \int_{\mathcal{S}} \left. \frac{d\mathcal{L}_f(s)}{ds} \right|_{s=\delta + j\omega} e^{(\delta + j\omega)t} d(\delta + j\omega) = \\
 &= (1 - \varepsilon t) \frac{f(t+0) + f(t-0)}{2} + \varepsilon(1 - \varepsilon t) \left[t \frac{f(t+0) + f(t-0)}{2} \right] =
 \end{aligned}$$

$$\begin{aligned}
 &= (1 - \epsilon t) (1 + \epsilon t) \frac{f(t+0) + f(t-0)}{2} = (1 - \epsilon^2 t^2) \frac{f(t+0) + f(t-0)}{2} = \\
 &= \frac{f(t+0) + f(t-0)}{2}.
 \end{aligned}$$

Remark. The existence of inversion formula (9), demonstrates the possibility to synthesise any signal in $\mathbb{L}_1^1$, by means of amplitude and phase modulated (4) like signals, representable symbolically on $\mathcal{S}$.

4. Let be a linear system with constant parameters on whose input a signal $s(t)$ representable symbolically on algebra $\mathcal{S}$ by element $\hat{w}$ is applied. The steady-state response of the signal is represented symbolically by element $\hat{w}\hat{H}$, where $\hat{H}$ (the transference of the sistem) is obtained by substituting, in the expression of Laplace transfer function $H(s)$, the variable s with the expression of the derivation operator $\hat{d}$ [1], [2].

The proposed integral transformation allows an original interpretation of the notion of „transference“ given by:

Theorem 4. *The transference of a linear system on algebra $\mathcal{S}$ is obtained by applying the transformation (5) to the weighting function of the system.*

Proof. The system response to excitation $s(t)$ is:

$$(10) \quad r(t) = h(t) * s(t),$$

where $h(t)$ is the weighting function of the considered system. Taking into account Theorem 2 and the above statements (10), it results:

$$\hat{H}(\hat{d}) = \hat{\mathbb{R}}_{h(t)}.$$

The following relation between transference and the Laplace transfer function of a system results from (7):

$$(11) \quad \hat{H}(\hat{d}) = H(s) \Big|_{s=\delta+j\omega} + \frac{\epsilon dH(s)}{ds} \Big|_{s=\delta+j\omega},$$

which is very useful in applications.

For instance, we shall establish the conditions which should obey the Laplace transfer function $H(s)$ of a system in order to obtain a pure amplitude modulated signal as response to a signal like:

$$(12) \quad s(t) = e^{\delta t}(at+b) \sin(\omega t + \varphi).$$

Signal (12) is representable symbolically on algebra $\mathcal{S}$ with the derivation operator $\hat{d} = \delta + \epsilon + j\omega$ by a special-type element. As in $\mathcal{S}$, the product between a certain element $\hat{a}$ and a special type element $\hat{b}$ is a special-type one, only if $\hat{a}$ is a special type element, the undistorsion condition imposes that the transference of the system should be a special type element.

From (11) and (4₁), it results:

$$(12) \quad \operatorname{Re} H(s) \operatorname{Im} \frac{dH(s)}{ds} = \operatorname{Im} H(s) \operatorname{Re} \frac{dH(s)}{ds}, \quad s = \delta + j\omega.$$

Condition (12) is equivalent with :

$$(12_1) \quad \operatorname{Im} \overline{H(s)} \cdot \frac{dH(s)}{ds} \Big|_{s=\delta+j\omega} = 0,$$

where — indicates the conjugation of complex numbers.

Condition (12₁) is less restrictive than the already known ones as it has to be satisfied in only one point.

Conditions may also be imposed in order to obtain a specified form for the response of the system or to reject the parasite phase modulation of the input signal.

New informations regarding the synthesis of such systems may be obtained by studying the compatibility among (12₁), like conditions and the realizability ones of the considered system.

Received, December 28, 1983

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TRANSFORMĂRI INTEGRALE PE ALGEERE (II)

(Rezumat)

Se propune, urmînd metodologia din [2], o transformare integrală pe algebra lui Sobrero. Se indică noi posibilități de aplicare a proprietăților demonstrate în studiul răspunsului sistemelor liniare la semnale modulate.

THE SYNTHESIS OF SOME TRIANGULAR WAVEFORMS

BY

VIOREL COSMA and D. ALEXANDRU

1. Introduction

The mathematical and experimental method of synthesis for a waveform $y(t)$, satisfying the conditions imposed by the Dirichlet's theorem is based upon the Fourier series expanding such as

$$(1) \quad y(t) = a_0 + \sum_{k=1}^{\infty} (a_k \cos k\omega t + b_k \sin k\omega t),$$

followed by an experimental generation of the functions resulted from the convenient grouping of the series terms (1), where the a_k and b_k coefficients as well as the angular frequency ω are arbitrarily imposed.

Passive components such as rectifying and Zenner diodes, linear resistors and active ones like negative impedance convertors and constant current sources are used for this purpose.

For the simplicity's sake the components implied in the synthesis of the resulting functions are assumed to be ideal ones.

2. Mathematical Synthesis of Triangular Waveforms

The aim of this work is the synthesis of the triangular waveforms $y(t)$ (s. Fig. 1) which are described by the equations

$$(2) \quad y(t) = \begin{cases} Y_m \frac{t}{t_r} & \text{for } t \in (0, t_r), \\ Y_m \frac{T-t}{T-t_r} & \text{for } t \in (t_r, T). \end{cases}$$

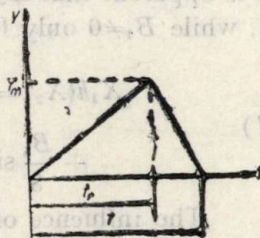


Fig. 1

In order to obtain a generalized method, a normalization of the parameters characterizing the signal to be synthesized is used:

$$(3) \quad \frac{Y}{Y_m} = \bar{y}, \quad \frac{t}{T} = \bar{t}, \quad \frac{t_r}{T} = \bar{t}_r,$$

where the normed restoring time $\bar{t}_r \in [0, 1]$.

By convenient grouping of the Fourier series' terms restricting ourselves to the 4th order terms [2] and expanding $\cos k2\pi\bar{t}$ and respectively $\sin k2\pi\bar{t}$ in polynoms of $\cos 2\pi\bar{t}$ and $\sin 2\pi\bar{t}$, the signal to be synthesized can be written as :

$$(4) \quad \bar{y}(\bar{t}) = f(X_2) + g(X_1) + X_1 h(X_2),$$

where :

$$(5) \quad X_2 = \cos 2\pi\bar{t}, \quad X_1 = \sin 2\pi\bar{t}$$

and

$$(6) \quad \begin{cases} f(X_2) = A_0 + A_1 X_2 + A_2 X_2^2 + A_3 X_2^3 + A_4 X_2^4, \\ g(X_1) = B_1 X_1 + B_3 X_1^3, \quad h(X_2) = B_2 X_2 + B_4 X_2^3. \end{cases}$$

The values of the A_k and B_k coefficients are given in the Table 1.

Table 1
Values of the A_k and B_k Coefficients

$\bar{t}_r$	0	$\frac{1}{4}$	$\frac{1}{2}$	$\frac{3}{4}$	1
A_0	0.5	0.63509	0.5	0.63509	0.5
A_1	0	-0.18013	-0.27017	-0.1813	0
A_2	0	-0.27019	0	-0.27019	0
A_3	0	-0.12008	-0.18014	-0.12008	0
A_4	0	0	0	0	0
B_1	0.6366	0.18014	0	-0.18014	-0.6366
B_2	0	0	0	0	0
B_3	0.4244	0.12008	0	-0.12008	0.4244
B_4	0.6366	0	0	0	-0.6386

It is apparent that B_2 is zero for all the particular values of the parameter $\bar{t}_r$, while $B_4 \neq 0$ only for the values $\bar{t}_r = 0$ and 1. In these cases :

$$(7) \quad \begin{aligned} X_1 h(X_2) &= B_4 X_1 X_2^3 = B_4 \sin 2\pi\bar{t} \cos^3 2\pi\bar{t} = \frac{B_4}{4} \sin 4\pi\bar{t} + \\ &+ \frac{B_4}{8} \sin 8\pi\bar{t} = \pm 0.1567 \sin 4\pi\bar{t} \pm 0.0796 \sin 8\pi\bar{t}. \end{aligned}$$

The influence of the term in $\sin 8\pi\bar{t}$ is practically negligible, thus in order to obtain some particular waveforms with $\bar{t}_r = 0$ or 1, a double frequency source is needed.

The graphic plots of $f(X_2)$ and $g(X_1)$ for all the proposed values of the $\bar{t}_r$ parameter are given in the Figs. 2 and 3.

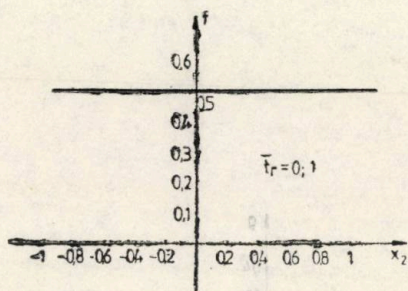


Fig. 2 a

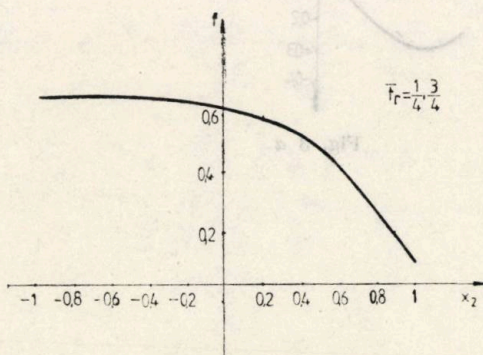


Fig. 2 b

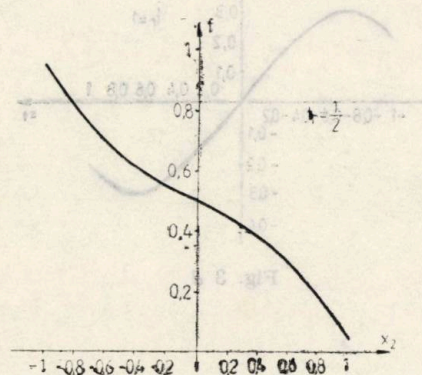


Fig. 2 c

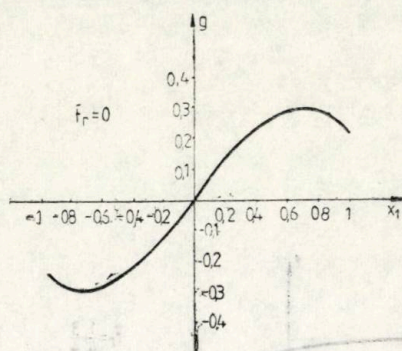


Fig. 3 a

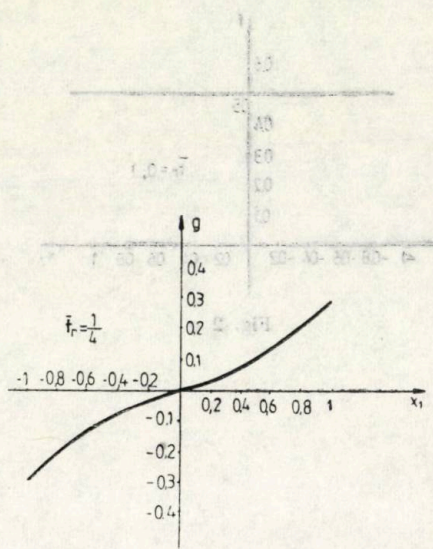


Fig. 3 b

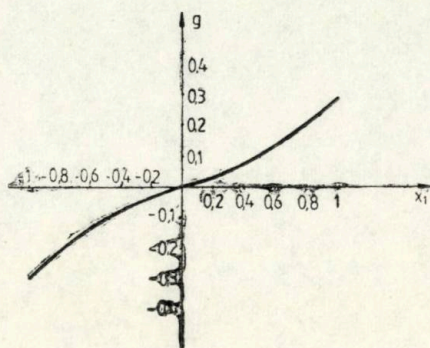


Fig. 3 c

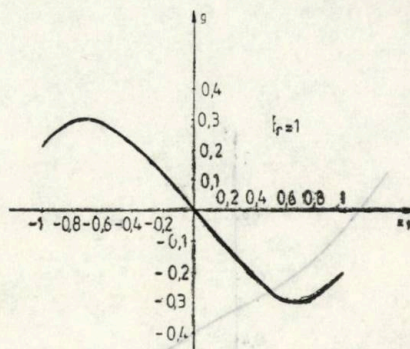


Fig. 3 d

3. Graphical and Experimental Synthesis of the $f(X_2)$ and $g(X_1)$ Functions

From the analysis of the non-linear functions $f(X_2)$ and $g(X_1)$ an odd symmetry in most cases is observed, which greatly simplifies the synthesis. The proposed non-linear resistors, with $i(v)$ characteristics representing the experimental realization of the non-linear functions f and g , can be obtained by means of concave resistors, INICs or VNICs and constant current sources.

3.1 Experimental Realization of the Non-linear Function $f(X_2)$

Experimental realization of the $f(X_2)$ function for the cases $\bar{t}_r=0$ and 1 implies a simple current source.

For the cases $\bar{t}_r=1/4$ and $3/4$ the non-linear resistor depicted in Fig. 4 a is proposed. Its characteristic is graphically synthesized in Fig. 4 b and represents the experimental realization of the non-linear function $f(X_2)$ from Fig. 2 b.

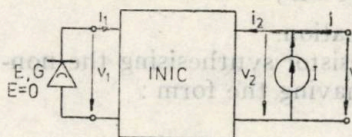


Fig. 4 a

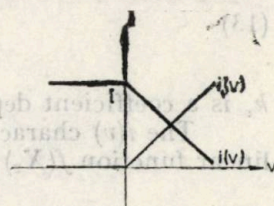


Fig. 4 b

Taking into account the equations describing the INIC and applying the Kirchhoff's theorems for the circuit depicted in Fig. 4 a we get,

$$(8) \quad \begin{bmatrix} v_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ k_I & 0 \end{bmatrix} \begin{bmatrix} i_1 \\ v_2 \end{bmatrix},$$

with

$$(9) \quad v = v_1 = v_2, \quad i_2 = k_I i_1, \quad i = i_2 + I,$$

where k_I is a coefficient determined by the INIC circuit.

The $i(v)$ characteristic of the synthesized non-linear resistor will be :

$$(10) \quad i = \begin{cases} I - k_I G v, & \text{for } v \geq 0, \\ I, & \text{for } v < 0. \end{cases}$$

For the experimental realization of the function $f(X_2)$ when $\bar{t}_r=1/2$ the non-linear resistor from Fig. 5 a is proposed, with the graphically synthesized characteristic depicted in fig. 5 b.

Starting from the Eq. (11) which describes the VNIC, applying the Kirchhoff's theorems (12) to the circuit represented in Fig. 5 a and having in mind the $i_3(v)$ characteristic of the concave resistors connected antiparallel (13), we obtain :

$$(11) \quad \begin{bmatrix} v_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} 0 & -k_v \\ -1 & 0 \end{bmatrix} \begin{bmatrix} i_1 \\ v_2 \end{bmatrix},$$

which

$$(12) \quad i_1 = I + i_3 + i_4, \quad v_2 = \frac{-v_1}{k_v},$$

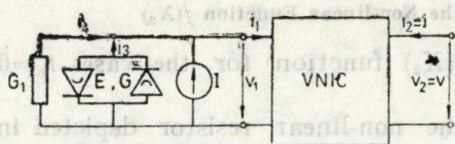


Fig. 5 a

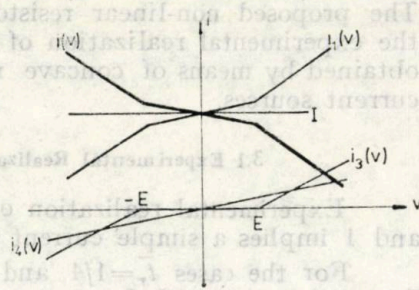


Fig. 5 b

$$(13) \quad i_3 = \begin{cases} 0 & \text{for } v_1 \in [-E, E], \\ -G(v_1 - E) & \text{for } v_1 > E, \\ -G(v_1 + E) & \text{for } v_1 < -E; \end{cases}$$

k_v is a coefficient depending on the VNIC realization.

The $i(v)$ characteristic of the non-linear resistor synthesising the non-linear function $f(X_2)$ from Fig. 2 c is obtained having the form :

$$(14) \quad i = \begin{cases} I - \frac{G_1}{k_v} v & \text{for } v \in [-E, E], \\ I + \frac{G}{k_v} E - \frac{G+G_1}{k_v} v & \text{for } v > E, \\ I - \frac{G}{k_v} E - \frac{G+G_1}{k_v} v & \text{for } v < -E. \end{cases}$$

3.2. Experimental Realization of the Non-linear Function $g(X_1)$

The graphical plots of the $g(X_1)$ function show an odd symmetry.

For the experimental realization of the $g(X_1)$ function in the case $\bar{i}_r = 1$ (Fig. 3 d), the non-linear resistor from Fig. 6 a is proposed with a N type characteristic as synthesized in [3], shown in Fig. 6 b.

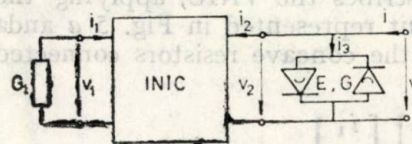


Fig. 6 a

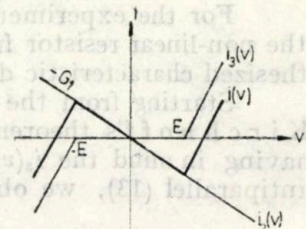


Fig. 6 b

Taking into account Eqs. (8), (13) and the Kirchhoff's theorems applied to the electrical circuit represented in Fig. 6 a, the equations of the proposed non-linear resistor are obtained, namely:

$$(15) \quad i = \begin{cases} -k_1 G_1 v & \text{for } v \in [-E, E], \\ -k_1 G_1 E + G(v - E) & \text{for } v > E, \\ k_1 G_1 E + G(v + E) & \text{for } v < -E. \end{cases}$$

The proposed non-linear resistor used for obtaining the $g(X_1)$ function in the case $\bar{t}_r = 0$ (Fig. 3 a), considering the symmetry of the characteristics is obtained by applying the former non-linear resistor from Fig. 6 a, to one of the VNIC's gates (Fig. 7 b).

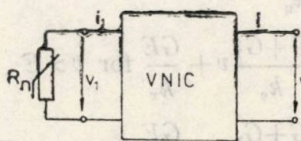


Fig. 7 a

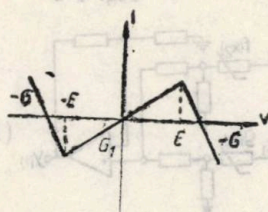


Fig. 7 b

Starting from Eqs. (15) and (11) the equations of the characteristic of the synthesized non-linear resistor are obtained:

$$(16) \quad i = \begin{cases} \frac{k_1}{k_e} G_1 v & \text{for } v \in [-E, E], \\ \frac{k_1}{k_e} G_1 E - \frac{G}{k_e} (v - E) & \text{for } v < E, \\ -\frac{k_1}{k_e} G_1 E - \frac{G}{k_e} (v + E) & \text{for } v < -E. \end{cases}$$

The experimental realization of the non-linear function $g(X_1)$ in the case $\bar{t}_r = 1/4$ was obtained with the non-linear resistor represented in Fig. 8, also found in the synthesis of the $f(X_2)$ function for $\bar{t}_r = 1/2$ (Fig. 5 a).

The equations of the characteristic of this non-linear resistor are obviously:

$$(17) \quad i = \begin{cases} G_1 v & \text{for } v \in [-E, E], \\ (G + G_1)v - GE & \text{for } v > E, \\ -(G + G_1)v + GE & \text{for } v < -E. \end{cases}$$

Taking into account the symmetry of the plots in Figs. 3 b and 3 c the synthesis of the non-linear resistor with the $i(v)$ characteristic which realizes the $g(X_1)$ function for the case $\bar{t}_r = 3/4$ involve the connection of the non-linear resistor which was synthesized in Fig. 8 to one of the VNIC's gates (s. Fig. 9 a). The characteristic is given in Fig. 9 b.

The equations describing the characteristic of the non-linear resistor which was synthesized are evidently:

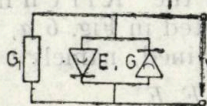


Fig. 8

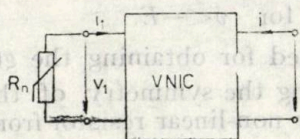


Fig. 9 a

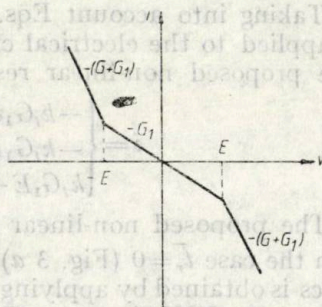


Fig. 9 b

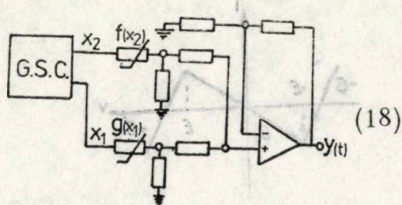


Fig. 10

$$i = \begin{cases} -\frac{G_1}{k_v} v & \text{for } v \in [E, -E], \\ -\frac{G+G_1}{k_v} v + \frac{GE}{k_v} & \text{for } v > E, \\ -\frac{G+G_1}{k_v} v - \frac{GE}{k_v} & \text{for } v < -E. \end{cases} \quad (18)$$

In order to obtain the desired triangular waveform, $y(t)$, the diagram given in Fig. 10 was used, where G.S.C. represents a quadrature sinewaves generator.

4. Conclusions

The mathematical and experimental method of synthesis of the periodic waveforms shown here is not a difficult one. The number of elements involved is quite small. The experimental results are in good agreement with the theoretical aspects presented above. In the cases $t_r = 1$ and 0, a double frequency source is used instead of a cosine source, as presented in Fig. 10.

Received, June 17, 1983

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SINTEZA UNOR FORME DE UNDĂ TRIUNGHIULARE

(Rezumat)

În vederea obținerii formelor de undă triunghiulare dorite se urmărește sinteza experimentală și grafică a funcțiilor rezultate din gruparea convenabilă a termenilor seriei Fourier în care s-au descompus respectivele forme de undă. Se folosesc în acest scop componente pasive ca: diode redresoare, diode Zenner, rezistoare liniare și componente active: convertitoare de impedanță negativă și generatoare de curent constant.

UPON THE DAMPING OF LATERAL OSCILLATIONS IN ELECTROMAGNETIC LEVITATION SYSTEMS USING U-SHAPED ELECTROMAGNETS AND RAILS

BY

SAVEL MITREA and LORIN CANTEMIR

1. Introduction

To the authors knowledge U-shaped electromagnets and rails are used in electromagnetic test vehicles as TRANSRAPID 04 in West Germany and HSST in Japan [1], [2]. It comes out from the papers concerning these vehicles that in order to ensure lateral guidance each „magnetic wheel“ consists of two electromagnets displaced laterally from the center line of the inverted U-shaped rail. The authors of the present paper proposes themselves to show, in a simple way, that this geometry allows the damping of lateral oscillations as well.

2. Outline of the System

The cross-section of rail-electromagnet system is shown in Fig. 1. It is assumed that four forces are applied to the electromagnet. The weight mg of electromagnet is compensated for by the lifting force F_{z_0} produced by a constant current i_0 corresponding to a constant applied voltage U_0 , in concordance with the given gap z_0 . In order to simplify the analysis it is assumed that any change in the lifting force caused by the time-varying terms of the applied voltage u cannot bring about a change in the gap due to a hypothetical mechanical system that allows the electromagnet to slide, without friction, laterally only. The other pair of forces consist of F_p and F_x . F_p is supposed to be a constant external force similar to the weight mg , that produces a constant lateral displacement p . F_x represents the magnetic restoring force in the x direction and varies with the displacement $p+x$.

3. The System Equations

The equations describing the system are the following:

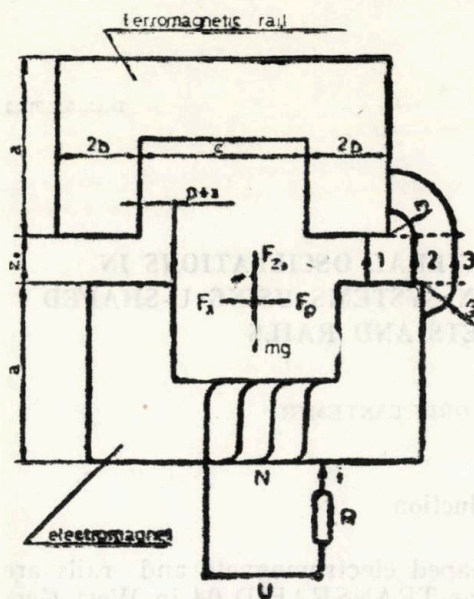


Fig. 1

The inductance per unit length of electromagnet is determined by assuming that the electromagnet is sufficiently long so as the magnetic field is laplacean. Also, magnetic voltage drop over the ferromagnetic path is ignored and magnetic field lines in free space are approximated as shown in Fig. 1. Under the above assumptions the inductance per unit length of electromagnet takes the form:

$$(3) \quad L = \mu_0 N^2 \left\{ \frac{2b - (p+x)}{2z_0} + \frac{2}{\pi} \ln \left[1 + \frac{\pi(p+x)}{2z_0} \right] + \frac{1}{\pi} \ln \left[1 + \frac{2\pi a}{2z_0 + \pi(p+x)} \right] \right\}.$$

In order that the above approximate expression not be too far from the exact one the pole-width $2b$ must be several times greater than the gap z_0 and the displacement $p+x$ must be positive and small in comparison with $2b$. Also, the dimensions c and a must be sufficiently large so as the approximation of magnetic field lines to hold, especially on the inner side of the geometry.

The restoring force in the x direction is obtained from the expression of magnetic field energy as follows:

$$(4) \quad F_x = \frac{i^2}{2} \cdot \frac{\partial L}{\partial x}.$$

In order that F_x vanish for $p+x=0$ we must let a become infinite after the differentiation of (3).

$$(1) \quad \begin{cases} Ri + \frac{d(Li)}{dt} = u, \\ m \frac{d^2x}{dt^2} = F_p + F_x. \end{cases}$$

They are derived by assuming that saturation, hysteresis, leakage flux and eddy currents are absent.

It is assumed that the applied voltage is of the form:

$$(2) \quad u = U_0 + k_v \frac{dx}{dt} + k_a \frac{d^2x}{dt^2},$$

where the first term U_0 is a constant voltage that ensures a current so as, at the given gap z_0 , the weight mg is compensated for by the force F_{z_0} , the second term is a voltage proportional to the velocity in the x direction and the third one is a voltage proportional to the acceleration in the same direction.

4. Stability

In order to investigate stability of the system for small perturbations in the vicinity of equilibrium point Eqs. (1) are written in normal form as follows:

$$(5) \quad \begin{cases} \frac{di}{dt} = L^{-1} \left[U_0 - Ri + \left(k_v - i \frac{\partial L}{\partial x} \right) \frac{dx}{dt} + k_a \frac{dv}{dt} \right] = f_1, \\ \frac{dx}{dt} = v = f_2, \quad \frac{dv}{dt} = m^{-1} (F_p + F_x) = f_3. \end{cases}$$

By equating to zero all the derivatives in (5) and making $x=0$, the equations of equilibrium point are obtained:

$$(6) \quad U_0 - Ri = 0, \quad F_p + F_x|_0 = 0.$$

The characteristic equation $f(\lambda) = 0$ is obtained under the form:

$$(7) \quad \begin{vmatrix} \left. \frac{\partial f_1}{\partial i} \right|_0 - \lambda & \left. \frac{\partial f_1}{\partial x} \right|_0 & \left. \frac{\partial f_1}{\partial v} \right|_0 \\ 0 & -\lambda & L \\ \left. \frac{\partial f_3}{\partial i} \right|_0 & \left. \frac{\partial f_3}{\partial x} \right|_0 & -\lambda \end{vmatrix} = 0,$$

where the partial derivatives take values corresponding to the equilibrium point. By expanding the determinant we obtain the third order algebraic equation:

$$(8) \quad a_0 \lambda^3 + a_1 \lambda^2 + a_2 \lambda + a_3 = 0,$$

whose coefficients have the form:

$$(9) \quad \begin{cases} a_0 = 1, & a_1 = \frac{R}{L_0} - \frac{k_a i_0}{m L_0} \cdot \left. \frac{\partial L}{\partial x} \right|_0, \\ a_2 = - \frac{k_v i_0}{m L_0} \cdot \left. \frac{\partial L}{\partial x} \right|_0 + \frac{i_0^2}{m L_0} \left(\left. \frac{\partial L}{\partial x} \right|_0 \right)^2 - \frac{i_0^2}{2m} \cdot \left. \frac{\partial^2 L}{\partial x^2} \right|_0, \\ a_3 = - \frac{R i_0^2}{2m L_0} \cdot \left. \frac{\partial^2 L}{\partial x^2} \right|_0. \end{cases}$$

According to Hurwitz criterion the system is stable in the vicinity of equilibrium point if the following conditions are fulfilled:

$$(10) \quad a_0, a_1, a_2, a_3 > 0 \quad \Delta = \begin{vmatrix} a_1 & a_0 \\ a_3 & a_2 \end{vmatrix} = a_1 a_2 - a_0 a_3 > 0.$$

Taking into account that the quantities

$$(11) \quad \left. \frac{\partial L}{\partial x} \right|_0 = \mu_0 N^2 \left(\frac{1}{2z_0 + \pi p} - \frac{1}{2z_0} \right), \quad \left. \frac{\partial^2 L}{\partial x^2} \right|_0 = - \frac{\mu_0 N^2 \pi}{(2z_0 + \pi p)^2}$$

are both negative, all the coefficients are positive even if velocity and acceleration feedback signals are absent. The second condition of stability is also fulfilled in the absence of feedback signals since in this case

$$(12) \quad \Delta = \frac{R i_0^2}{m L_0^2} \left(\left. \frac{\partial L}{\partial x} \right|_0 \right)^2 > 0,$$

provided that the electromagnet is displaced by a constant force, that is $p \neq 0$. Nevertheless, velocity and acceleration feedback signals are useful for the manipulation of the roots of characteristic equation. The coefficients of characteristic equation can be written in terms of its roots as follows:

$$(13) \quad a_1 = \lambda_1 + \lambda_2 + \lambda_3, \quad a_2 = \lambda_1\lambda_2 + \lambda_1\lambda_3 + \lambda_2\lambda_3, \quad a_3 = \lambda_1\lambda_2\lambda_3.$$

We may find it appropriate that the roots be real and equal one to another; then

$$(14) \quad \lambda_1 = \lambda_2 = \lambda_3 = \sqrt[3]{a_3},$$

where a_3 is given by (9). In this case the coefficients a_1 and a_2 can be adjusted easily to the right values by aid of k_a and k_v respectively.

One may find that the transient is too slow or too fast due to the value the coefficient a_3 takes according to the quantities it is consisted of. It easily seen that by applying a current feedback of the form $k_i i$, an additional term appear along with R so that a_3 can be adjusted to our convenience as well. It should be noted that by using a current feedback the constant voltage U_0 is to be adjusted accordingly so as the current i_0 from (6₁) be the same; otherwise the lifting force F_{20} will not be equal to the weight mg any more and the gap z_0 will not preserve its prescribed value. Improved transients, measured experimentally, are reported in [3] by using current feedback and the corresponding increase of d.c. source voltage supplying the electromagnet.

5. Conclusions

The damping of lateral oscillations of the electromagnet is possible only if a displaced geometry is used. The damping can be natural if no feedback is used. In order to adjust the damping to prescribed requirements, velocity, acceleration and current feedback signals may be used.

Received, May 26, 1983

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ASUPRA AMORTIZĂRII OSCILAȚIILOR LATERALE ÎN SISTEMLILE DE LEVITAȚIE ELECTROMAGNETICĂ UTILIZÂND ȘINE ȘI ELECTROMAGNEȚI DIN PROFIL U

(Rezumat)

Se arată că amortizarea oscilațiilor laterale este posibilă numai dacă acestea au loc în jurul unei poziții deplasate a electromagnetului față de axa de simetrie longitudinală a șinei de ghidare. Amortizarea poate fi naturală dacă nu se utilizează semnale de reacție inversă. În scopul realizării unei amortizări în conformitate cu cerințe prescrise se pot utiliza reacții inverse de viteză, accelerație și curent.

METODĂ ȘI APARAT PENTRU MĂSURAREA DIRECTĂ A PARAMETRILOR BOBINELOR LINIARE

DE

D. BĂRBULESCU și ADRIAN CREȚU

1. Introducere

În stadiul actual s-a dezvoltat o diversitate mare de metode și aparate destinate măsurării parametrilor R , L , C ai circuitelor de curent alternativ [5].

Metoda și aparatul propuse și studiate în prezenta lucrare, continuând cercetări anterioare ale autorilor [1] ... [3], evidențiază noi posibilități de utilizare a redresoarelor sensibile la fază în scopul măsurării directe a parametrilor bobinelor liniare.

2. Schema structurală de măsurare

Metoda propusă, analizată în acest paragraf, permite măsurarea directă a parametrilor bobinelor liniare prin determinarea componentelor activă și reactivă ale tensiunii de la bornele bobinei de măsurat cu ajutorul unui redresor sensibil la fază (RSF), asociat cu un defazor și un instrument de măsură magnetoelectric.

Schema de principiu este reprezentată în fig. 1, iar diagrama fazorială a curenților și tensiunilor din circuitul de măsurare, în fig. 2.

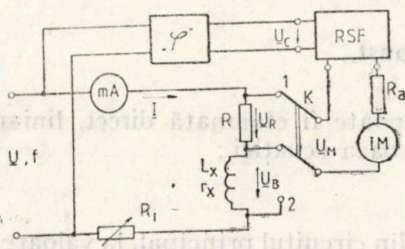


Fig. 1

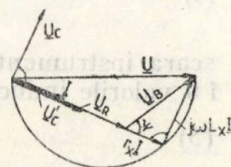


Fig. 2

Circuitul principal, constituit din bobina liniară L_x , r_x , rezistorul liniar fix R , rezistorul variabil R_i și miliampermetrul mA , este alimentat de la o sursă de t.e.m. sinusoidală cu valoarea efectivă U și frecvența f , constante.

Pentru măsurarea directă a parametrilor L_x , r_x se utilizează un redresor sensibil la fază, electronic, RSF, asociat cu defazorul φ și instrumentul magnetoelectric IM.

Măsurarea inductanței L_x se efectuează în două etape:

a) Se conectează comutatorul K pe poziția 1, aplicînd circuitului de măsurare al RSF semnalul $\underline{U}_M = \underline{U}_R = RI$ de la bornele rezistorului fix R . Se reglează faza φ a semnalului u_c aplicat circuitului de comandă al RSF, pînă cînd deviația α a instrumentului magnetoelectric IM devine egală cu zero.

Deviația α se exprimă prin relația [4]:

$$(1) \quad \alpha = S_I \frac{U_M}{2k_f R_a} \cos(\widehat{\underline{U}_M, \underline{U}_C}) = S_I \frac{U_R}{2k_f R_a} \cos(\widehat{\underline{U}_R, \underline{U}_C}),$$

în care: S_I reprezintă sensibilitatea la curent a instrumentului IM, $k_f = 1,11$ — factorul de formă al semnalului sinusoidal iar R_a — rezistența totală a circuitului de măsurare al RSF (în R_a se includ și rezistența de ieșire a RSF și rezistența internă a instrumentului IM).

Din (1) rezultă că la anularea deviației α , semnalele $\underline{U}_R$ și $\underline{U}_C$ sînt aduse în cuadratură.

b) Se trece comutatorul K pe poziția 2, caz în care circuitului de măsurare al RSF i se aplică semnalul:

$$(2) \quad \underline{U}_M = \underline{U}_B = (r_x \oplus j\omega L_x)I.$$

În această situație, ținînd cont de (1) și (2), deviația α a instrumentului IM devine:

$$(3) \quad \alpha = S_I \frac{U_B}{2k_f R_a} \cos(\widehat{\underline{U}_B, \underline{U}_C}) = S_I \frac{U_B}{2k_f R_a} \cos\left(\frac{\pi}{2} - \psi\right) = S_I \frac{\omega L_x I}{2k_f R_a},$$

în care $\psi = (\widehat{\underline{U}_B, I})$ (v. fig. 2).

Dacă se notează constanta aparatului cu:

$$(4) \quad k_L = \frac{2k_f R_a}{S_I \omega I} = \text{const.},$$

scara instrumentului magnetoelectric IM poate fi etalonată direct, liniar, în valorile inductanței L_x de măsurat, pe baza ecuației:

$$(5) \quad L_x = k_L \alpha.$$

Mentținerea intensității curentului, I , din circuitul principal, la valoarea necesară, rezultată pe baza limitelor domeniului de măsurare, se realizează cu rezistorul variabil R_t .

Măsurarea rezistenței r_x a bobinei se efectuează astfel:

a) Cu comutatorul K pe poziția 1 se reglează faza φ a semnalului de comandă al RSF pînă cînd deviația α a instrumentului IM devine maximă, respectiv:

$$(6) \quad \alpha = S_I \frac{U_R}{2k_f R_a} \cos(\underline{U}_R, \underline{U}'_C) = S_I \frac{U_R}{2k_f R_a}.$$

În această situație semnalul de comandă $\underline{U}'_C$ al RSF are aceeași valoare efectivă ca și semnalul $\underline{U}_C$ din cazul măsurării inductanței, dar este în cuadratură cu acesta, fiind în fază cu semnalul I (fig. 2).

b) Se trece comutatorul K pe poziția 2, deviația α a instrumentului IM devenind în acest caz proporțională cu componenta activă a semnalului $\underline{U}_B$, respectiv :

$$(7) \quad \alpha = S_I \frac{U_B}{2k_f R_a} \cos(\underline{U}_B, \underline{U}'_C) = S_I \frac{U_B}{2k_f R_a} \cos \psi = S_I \frac{r_x I}{2k_f R_a}.$$

Notînd :

$$(8) \quad k_R = \frac{2k_f R_a}{S_I I} = \text{const.},$$

constanta aparatului în noua situație, scara instrumentului IM poate fi etalonată direct, de asemenea liniar, în valorile rezistenței r_x a bobinei de măsurat, pe baza ecuației :

$$(9) \quad r_x = K_R \alpha.$$

3. Aparat pentru măsurarea directă a parametrilor bobinelor liniare (henrymetru)

Procedeul descris mai sus stă la baza funcționării henrymetrului a cărui schemă electrică este reprezentată în fig. 3.

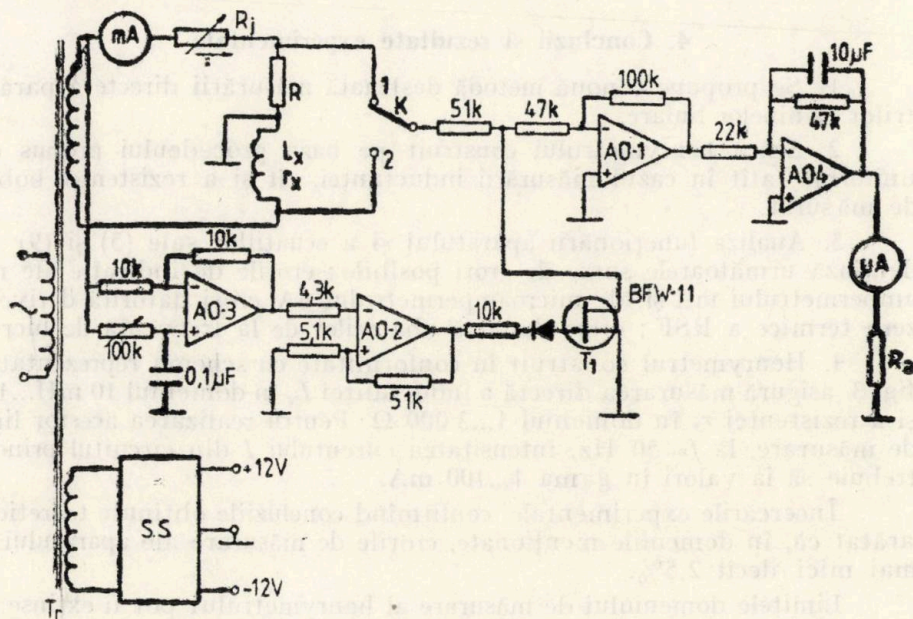


Fig. 3

Circuitul principal al aparatului, constituit din aceleași elemente ca și cel din fig. 1, este alimentat de la rețeaua de curent alternativ de frecvență industrială ($f=50$ Hz), prin intermediul transformatorului coborîtor T_r (220/10 ; 20 V).

Măsurarea directă a componentelor semnalului U_B , direct proporționale cu parametrii L_x , r_x ai bobinei de măsurat, se efectuează cu un redresor-amplificator sensibil la fază, de concepție proprie [4], asociat cu un defazor electronic și cu microampermetrul μA , de 100 μA .

Elementele principale ale RSF sînt : amplificatorul operațional AO-1 și un tranzistor cu efect de cîmp T_1 , comandat pe grilă cu semnalul dreptunghiular obținut la ieșirea unui trigger Schmitt, realizat cu amplificatorul operațional AO-2 [7]. Reglarea fazei φ a semnalului de comandă U_c , în limitele $[0, 180^\circ]$, se efectuează prin intermediul unui defazor construit cu amplificatorul operațional AO-3, la intrarea căruia se aplică un semnal sinusoidal simfazic cu curentul I din circuitul principal.

Semnalul de ieșire al RSF este aplicat la intrarea filtrului operațional, realizat cu amplificatorul operațional AO-4, în al cărui circuit de ieșire este conectat microampermetrul μA .

Alimentarea amplificatoarelor operaționale AO-1, ..., AO-4, de tipul 741, se realizează cu tensiunea continuă ± 12 V obținută prin redresarea și stabilizarea tensiunii de la bornele celui de al doilea secundar al transformatorului T_r (sursa S.S.).

Constantele k_L și k_R ale henrymetrului din fig. 3 rezultă prin multiplicarea expresiilor (4), respectiv (8), cu factorii de amplificare ai RSF și filtrului operațional.

4. Concluzii și rezultate experimentale

1. Se propune o nouă metodă destinată măsurării directe a parametrilor bobinelor liniare.

2. Scara henrymetrului construit pe baza procedurii propus este uniformă, atît în cazul măsurării inductanței, cît și a rezistenței bobinei de măsurat.

3. Analiza funcționării aparatului și a ecuațiilor sale (5) și (9) evidențiază următoarele surse de erori posibile : erorile de indicație ale miliampermetrului mA și ale microampermetrului μA ; erori datorită derivatei de zero termice a RSF ; erori datorită abaterilor de la frecvența de lucru f .

4. Henrymetrul construit în conformitate cu schema reprezentată în fig. 3 asigură măsurarea directă a inductanței L_x în domeniul 10 mH...10 H și a rezistenței r_x în domeniul 1...3 000 Ω . Pentru realizarea acestor limite de măsurare, la $f=50$ Hz, intensitatea curentului I din circuitul principal trebuie să ia valori în gama 1...100 mA.

Încercările experimentale, confirmînd concluziile obținute teoretic, au arătat că, în domeniile menționate, erorile de măsurare ale aparatului sînt mai mici decît 2,5%.

Limitele domeniului de măsurare al henrymetrului pot fi extinse prin utilizarea unor frecvențe de lucru mai mari decît frecvența industrială.

5. Comparativ cu alte metode și aparate cunoscute, majoritatea cu măsurare indirectă, procedeul propus oferă posibilitatea construirii unor henrymetre cu scara liniară, relativ simple și, mai ales, lipsite de elemente de circuit etalon, scumpe. Aceste aparate pot fi realizate numai cu componente produse în țară.

Primită la 9 iulie 1983

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MÉTHODE ET APPAREIL POUR LA MESURE DIRECTE DES PARAMÈTRES DES BOBINES LINÉAIRES

(Résumé)

On propose et on étudie un nouveau procédé pour la mesure directe des paramètres des bobines linéaires, basé sur l'utilisation d'un redresseur-amplificateur sensible à la phase, conçu par les auteurs.

Le henrymètre construit, basé sur ce procédé, permet la mesure directe des inductivités et des résistances des bobines linéaires, avec des erreurs moindres que 2,5%, les deux échelles de l'appareil en étant uniformes.

CURRENT TRANSFORMER

BY

ELENA CRETU, M. LUCANU and ADRIAN CRETU

1. Introduction

In the specialized literature there are presented a great number of electronic compensation diagrams of the remarkable performance current transformers errors due to both the very diminished values of the errors and to the savings of magnetic material they accomplish.

As the current transformers compensated by means of electronic methods work, practically, in a current comparator regime, so, at a null flux, the secondary current is supplied in such a case, either by the correction electronic amplifier [1] or by an auxiliary current transformer [2], [3], the electronic amplifier generating but the magnetizing current corresponding to the error of the compensated current transformer.

As the electronic procedures have been improved, the idea of using an auxiliary transformer has been given up, the electronic compensated current transformer having two cores [4], [5], one working at null flux, corresponding to the compensated transformer and another one in which the magnetic flux is concentrated and corresponding to the auxiliary transformer.

In what follows it is presented an electronic compensation method of the errors, that can be applied, without constructional changes, to the two-stage-current-transformers [6] and, adding a detection winding to the two (or more) secondary windings transformers [7] if these transformers have their secondaries wound on different cores.

2. Schematic Diagram

In Fig. 1 it is presented the principle diagram of the two secondary windings electronic compensated current transformer where the secondary SI together with the primary N_1 make up the indicator transformer TI , while the secondary SA and the primary N_1 form the auxiliary transformer TA [7].

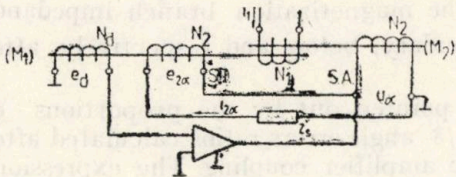


Fig. 1

At the electronic amplifier input it is applied the induced voltage in the N_D turns detection winding placed

on the same core with *SI* indicator secondary while the output voltage (the U_α correcting voltage) is applied at the *SA* auxiliary secondary terminals. Thus, the electronic amplifier corrects the secondary current value from i_2 to $i_{2\alpha}$ determining at the same time a magnetic flux diminishing in the *SI* secondary core and its concentration in the one of *SA* secondary. As a consequence, the *SA* auxiliary secondary corresponding to the *TA* transformer of a greater power and more reduced precision, maintains the secondary current while the *SI* secondary, corresponding to *TI* indicator transformer, compensated with the correction amplifier works with high precision up to the power corresponding to *TA* transformer.

The magnetic flux diminution in the *SI* secondary core determines the modification of the operating point on the magnetization characteristic of this core from the value B_m of the magnetic induction amplitude, to $B_{m\alpha}$ [8]:

$$(1) \quad B_{m\alpha} = \frac{B_m}{AN_D/N_2}.$$

Relation (1) shows that after the electronic amplifier coupling, a diminution of B_m magnetic induction amplitude takes place, and so, of the magnetic flux Φ_m amplitude in the core of the *SI* indicator secondary; the higher the values of the detection winding N_D number of turns and of the electronic amplifier A gain, the more obvious this diminution will be. For an A voltage gain, theoretically infinite, the *TI* current transformer operates in a null flux, in a current comparator regime with minimal errors. At the same time, due to the magnetic flux vanishing in the *SI* secondary core, the dimensions of this core can be much diminished.

The operating point changing on the magnetization characteristic leads to the modification of the idle running ampere turns from the $N_1 I_0$ value to $N_1 I_{0\alpha}$, of the magnetic permeability from μ to μ_α as well as of the loss angle from α to α_α . The ratio as well as angle errors, accordingly, change their values from ε_i to $\varepsilon_{i\alpha}$ and from δ_i to $\delta_{i\alpha}$, respectively. The complex error changes from the value [9]:

$$(2) \quad \varepsilon_i + j\delta_i = \frac{Z_2 + Z_s}{Z'_0},$$

to the value:

$$(3) \quad \varepsilon_{i\alpha} + j\delta_{i\alpha} = \frac{Z_2 + Z_s}{Z'_{0\alpha}} \cdot \frac{1}{(1+A) N_D/N_2},$$

where Z_2 represents the *SI* secondary winding impedance, Z_s — the load impedance, while Z'_0 , $Z'_{0\alpha}$ represents the magnetization branch impedance of *TI* transformer related to the secondary, before and respectively, after the electronic amplifier coupling.

The compensation efficiency is pointed out by the proportions of the $\varepsilon_{i\alpha}/\varepsilon_i$ current errors and of the $\delta_{i\alpha}/\delta_i$ angle errors ratios calculated after and respectively before the electronic amplifier coupling. The expressions of these proportions have the form [8]:

$$(4) \quad \frac{\varepsilon_{i\alpha}}{\varepsilon_i} = \frac{\cos \alpha}{\cos \alpha_x} \cdot \frac{\mu}{\mu_x} \cdot \frac{1}{AN_D/N_2} \cdot \frac{\sin(\alpha_x + \psi_{2\alpha})}{\sin(\alpha + \psi_2)}$$

and respectively :

$$(5) \quad \frac{\delta_{i\alpha}}{\delta_i} = \frac{\cos \alpha}{\cos \alpha_x} \cdot \frac{\mu}{\mu_x} \cdot \frac{1}{AN_D/N_2} \cdot \frac{\cos(\alpha_x + \psi_{2\alpha})}{\cos(\alpha + \psi_2)},$$

where ψ_2 represents the angle of phase displacement between $\underline{E}_2$ induced voltage and $\underline{I}_2$ secondary current before the electronic amplifier coupling, while $\psi_{2\alpha}$ represents the angle of phase displacement between $\underline{E}_{2\alpha}$ voltage induced in the SI secondary winding and $\underline{I}_{2\alpha}$ secondary current after the coupling of the correction electronic device.

The expressions (4) and (5) show that the higher the values of both A gain of the electronic device and μ_x permeability corresponding to the new operating point on the magnetization characteristic are and the smaller the loss angle is, the higher the compensation efficiency it will be.

As the TI electronic compensated transformer operates in a current comparator regime it follows that the greater the μ_x initial permeability of the SI secondary core and the smaller its iron losses are the higher the compensation efficiency it will be. If the core of SI secondary has not a high initial permeability, the efficiency of the compensation can be much improved by means of preliminary magnetization [10].

The application of preliminary magnetization is possible if SI secondary as well as N_D detection winding wind two superposed identical toroidal cores, M'_1 and M''_1 (Fig. 2), which substitute the initial core of SI secondary.

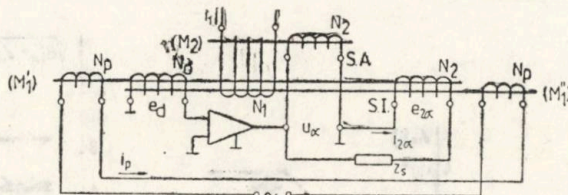


Fig. 2

The two N_D preliminary magnetization windings with the same number of turns, one wound on M'_1 core and the other one on M''_1 core, are connected in opposition in such a way as the Φ_p resultant flux created by i_p current to be null in order that it should not influence the values of both $i_{2\alpha}$ secondary current and of the e_d voltage, as well.

3. Experimental Results

Experimentally, it has been studied comparatively the errors at two T_1 and T_2 two secondary windings electronic compensated current transformers by means of the procedure pointed in Fig. 1.

The two transformers differ from each other due to the quality of the cores of SI indicator secondaries. Thus, while the SA auxiliary secondary cores of the two transformers as well as the indicator secondary core

of T_1 transformer are usually used in the achievement of CIS 0.5 transformers, the indicator core of T_2 transformer originates in a NORMA current transformer with 0.2 precision classes.

Initially, it has been determined the errors for the load impedance $Z=Z_n$ at a power factor of the load $\cos \varphi_s=1$.

It has been ascertained that :

a) For currents comprised between $(0.2...1.2) I_{1n}$ at the industrial frequency, the errors of the T_1 transformer are lower than 50 ppm while those of T_2 transformer do not exceed 10 ppm.

b) The errors of T_1 transformer decrease below 10 ppm by means of preliminary magnetization application.

These results have been obtained for a detection winding having the same number of turns like the secondary one ($N_D=N_2$).

The increase of the detection winding number of turns has not led to a more evident decrease of the errors.

For T_2 transformer it has been studied the errors variation function of voltage A gain of the correction electronic amplifier. In Figs. 3 and 4 it has been plotted the variation curves of both current errors $\epsilon_{i\alpha}$ and of angle errors $\delta_{i\alpha}$ as well, while in Fig. 5 it has been drawn the complex error modulus variation $\sqrt{\epsilon_{i\alpha}^2 + \delta_{i\alpha}^2}$ function of I_1 primary current at the nominal load impedance for different values of voltage A gain of the correction electronic amplifier. It is ascertained that the increase over a certain value ($A=2\ 000$) of the voltage gain is not justified any longer because it does not lead to the compensation efficiency increase. For instance, it is ascertained (Fig. 5) that the same values for the complex error modulus are obtained both for $A=2\ 000$ and $A=3\ 000$. It may also be noticed that at the voltage

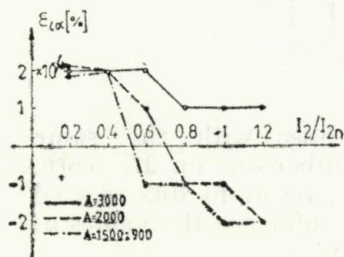


Fig. 3

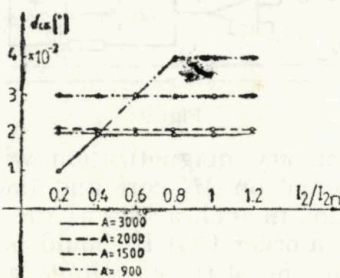


Fig. 4

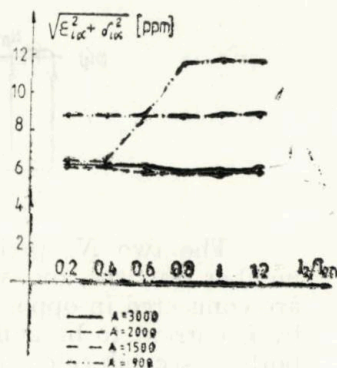


Fig. 5

gain increase, it does not take place an evident modification of the ratio error which maintains values between $\pm 0.0002\%$ but mainly an angle error diminution from $0.04'$ to $0.02'$.

Finally, it has been modified the load impedance between Z_n and $3Z_n$ and it has been ascertained that for currents lower than I_{1n} the errors remain very low, under 10 ppm.

4. Conclusions

1. A new electronic compensation method of the current transformer errors has been studied theoretically and experimentally, method which can be applied, without constructional changes, to two stage current transformers and, by adding a detection winding, to two secondary windings transformers, if these transformers have the secondaries wound on different cores.

2. It has been pointed out that the compensation efficiency depends on both the correction electronic amplifier parameters and on the core qualities.

3. When the core SI indicator secondary has not a great initial permeability, the performances of the method can be much improved by means of preliminary magnetization.

4. On an experimental device at the power frequency, errors below 10 ppm have been obtained both for nominal load impedance as well as for a multiple of it.

Received, October 25, 1982

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TRANSFORMATOR DE CURENT CU DOUĂ SECUNDARE, COMPENSAT ELECTRONIC

(Rezumat)

Se prezintă o schemă de compensare electronică a erorilor care poate fi aplicată, fără modificări constructive, transformatoarelor de curent în două trepte și, prin adăugarea unei înfășurări de detecție, transformatoarelor cu două (sau mai multe) înfășurări secundare, dacă aceste transformatoare au secundarele bobinate pe miezuri diferite.

Pe un montaj experimental, la frecvența industrială, s-au obținut erori sub 10 ppm atît pentru impedența de sarcină nominală cit și pentru un multiplu al acesteia.

4. Conclusions

A new electronic compensation method on the current transformer error has been studied theoretically and experimentally. The method can be applied without compensation changes to two-stage current transformers and by adding a detector winding to two secondary windings transformers. If these transformers have the secondary wound on different cores.

It has been pointed out that the compensation efficiency depends on both the correction electronic amplifier parameters and on the core properties.

When the core SV indicator secondary has not a great initial permeability, the performance of the method can be much improved by means of preliminary magnetization.

In an experimental device at the power frequency, errors below 10 ppm have been obtained both for nominal load impedances as well as for a multiple of it.

Received March 22, 1974
 Revised May 10, 1974
 Accepted June 10, 1974

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UDC 621.372.6.01
 THE THEORY OF THE TRANSFORMER

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PATHOLOGICAL PHENOMENA IN THE KIRCHHOFF INTERCONNECTED HYPERSYSTEMS

BY
M. VOICU

1. Introduction

Continuing the analysis of the properties of the Kirchhoff interconnectable systems and of the Kirchhoff interconnected hypersystems from [1], in this paper one identifies some pathological phenomena in the Kirchhoff interconnected hypersystems, of which symptoms are identically or analogically to there produced in case of the instability, and one formulate the conditions which insure the rejection of such undesirable phenomena. These new results are briefly presented and they suppose that the reader knows the results and the notations from [1].

The Kirchhoff interconnected hypersystem structurally contains a multiple feedback realised in condition which enabled systems S_i , ($i=1, \dots, n$), and S_β to have the same output y_β . In order of Theorem 1, [1], the systems S_i , ($i=1, \dots, n$), naturally realise also the functioning

$$(1) \quad \begin{cases} Y_{\alpha i}(s) = [G_{\alpha i}(s) - G_{\alpha \beta i}(s) G_{\beta i}^R(s) G_{\beta \alpha i}(s)] U_{\alpha i}(s) + \\ \quad + G_{\alpha \beta i}(s) G_{\beta i}^R(s) Y_\beta(s) + G_{\alpha \beta i}(s) [I_{m\beta} - G_{\beta i}^R(s) G_{\beta i}(s)] Z_i(s), \\ (2) \quad \begin{cases} U_{\beta i}(s) = -G_{\beta i}^R(s) G_{\beta \alpha i}(s) U_{\alpha i}(s) + G_{\beta i}^R(s) Y_\beta(s) + \\ \quad + [I_{m\beta} - G_{\beta i}^R(s) G_{\beta i}(s)] Z_i(s), \quad (i=1, \dots, n), \end{cases} \end{cases}$$

which follows from the Kirchhoff interconnectability (partially reversibility) of the systems S_i , ($i=1, \dots, n$).

The Kirchhoff interconnected hypersystem S , consisting of the systems S_i , ($i=1, \dots, n$), and S_β , has for the output $Y_\beta(s)$ the following form:

$$(3) \quad (S) \quad Y_\beta(s) = G_{\beta\beta}^{-1}(s) G_\beta(s) \left\{ \sum_{i=1}^n G_{\beta i}^R(s) G_{\beta \alpha i}(s) U_{\alpha i}(s) + \right. \\ \left. + \sum_{i=1}^n [G_{\beta i}^R(s) G_{\beta i}(s) - I_{m\beta}] Z_i(s) \right\}.$$

2. Pathological Phenomena

According to the structural localization of the pathological phenomena, they may be divided in three groups:

a) Kirchhoff uninterconnectability pathological phenomena, generated from the fact that at least one of the component systems is not Kirchhoff interconnectable, respective the condition

$$(4) \quad \text{rank } G_{\beta i}(s) = p_{\beta}$$

is not satisfied.

Such a system doesn't realise naturally the functioning (1), (2), whatever Y_{β} , $G_{\beta \alpha i} U_{\alpha i}$ may be. Consequently its connection into a Kirchhoff interconnected hypersystem is always followed from its rejection owing to the incompatibility, respective because the consistency condition (28) from [1] cannot be satisfied for every pair Y_{β} , $G_{\beta \alpha i} U_{\alpha i}$.

b) Arbitrariness pathological phenomena, generated from the fact that the partially reversing systems S'_i , ($i=1, \dots, n$), which insure the normal functioning of the Kirchhoff interconnection, are arbitrarily disturbed from Z_i , ($i=1, \dots, n$).

The disturbances Z_i , ($i=1, \dots, n$), are internally located and consequently they influence by an uncontrollable manner the output Y_{β} .

c) Singularity pathological phenomena, generated from the fact that the Kirchhoff interconnected hypersystem does not satisfy the condition

$$(5) \quad \det G_{\beta \gamma}(s) \neq 0.$$

Under these circumstances the functioning (3) is not possible, what implies the partial or total disintegration of the hypersystem because the consistency condition associated to the equation (40) from [1], in case $\det G_{\beta \gamma}(s) \equiv 0$, cannot be satisfied for every pair $U_{\alpha i}$, Z_i , ($i=1, \dots, n$).

Certainly, we are interested to know the conditions for which such pathological phenomena, of which symptoms are identically or analogically to there produced by the instability, can be naturally rejected. In connection with this question one can easily formulate the following assertion.

Theorem. A Kirchhoff interconnected hypersystem S is not affected from the pathological phenomena a), b), c) if and only if the component systems are Kirchhoff interconnectable, the hypersystem has non-singular multiple feedback with respect to the output Y_{β} and

$$(6) \quad m_{\beta} = p_{\beta}.$$

Proof.

Sufficiency. If $G_{\beta i}$, ($i=1, \dots, n$), and $G_{\beta \gamma}$ are quadratic and non-singular, then the hypersystem S is described relative to the Kirchhoff interconnection by the equation (47) from [1]. Under these circumstances S is not affected from the pathological phenomena a), b), c).

Necessity. If S is not affected from the pathological phenomena a), b), c), then $\text{rank } G_{\beta \gamma}(s) = p_{\beta}$ and, according to (3), we have $I_{m_{\beta}} -$

$-G_{\beta i}^R(s) G_{\beta i}(s) = 0$, ($i=1, \dots, n$). The latter condition implies that $G_{\beta i}^R(s)$, ($i=1, \dots, n$), are also left inverses of $G_{\beta i}(s)$, ($i=1, \dots, n$), what is true only if $m_{\beta} = p_{\beta}$.

Notice that although the hypersystem (47) from [1] is not affected from the pathological phenomena a), b), c) in order of Theorem above, one cannot assert anything about the output Y_{β} controllability with respect to each input $U_{\alpha i}$, ($i=1, \dots, n$). By completing Theorem above with the condition of the output Y_{β} controllability of the system S_{β} with respect to U_{β} , what also insures the output Y_{β} controllability of the system (47) from [1] with respect to each $U_{\alpha i}$, ($i=1, \dots, n$), one reaches in fact to Theorem 10 from [1].

3. Conclusions

The pathological phenomena of the Kirchhoff interconnected hypersystems are generated from the structure of the component systems and also from the Kirchhoff interconnection itself. It is to expect that the symptoms of these pathological phenomena are identically or analogically to there produced by the instability. Here we consider some phenomena which are possible in the hypersystems consisting of parallel-connected synchronous generators and which can be explained, certainly by means of a linearized model, by starting from the possible pathological phenomena in a Kirchhoff interconnected hypersystem.

Received, February 12, 1983

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FENOMENE PATOLOGICE ÎN HIPERSISTEMELE KIRCHHOFF INTERCONECTATE

(Rezumat)

Continuând analiza din [1] a proprietăților sistemelor Kirchhoff interconectabile și ale hipersistemelor Kirchhoff interconectate, se studiază unele fenomene patologice identice sau analoge, din punctul de vedere al simptomelor, cu fenomenele de instabilitate, proprii hipersistemelor Kirchhoff interconectate și se formulează condițiile în care acestea pot fi eliminate din funcționarea normală a lor.

Fig. 1

Le premier pas dans le processus d'analyse d'un schéma logique combinational ou séquentiel c'est la détermination de l'expression logique afférente à la variable ou aux variables de sortie. Dans ce but les variables de réaction, attachées à celles de sortie, sont désignées par les mêmes lettres, mais minuscules cette fois. De même aux entrées de la porte générant la variable de sortie doivent être présentes, d'une manière directe ou indirecte, les informations concernant toutes les variables primaires.

Dans notre exemple on désigne par q et $\bar{q}$ les variables de réaction (secondaires) et par Y_1, Y_2, Y_3 les sorties de trois portes logiques auxiliaires qui interviennent dans la structure du circuit.

En inspectant le circuit, il en résulte :

$$\begin{aligned} Y_1 &= \overline{qJT}; & Y_2 &= \overline{qKT}; & Y_3 &= \overline{q\bar{R}Y_2} = \overline{q\bar{R} \cdot \overline{qKT}}; \\ (1) \quad Q &= \overline{sY_1Y_3} = \overline{s \cdot \overline{qJT} \cdot \overline{q\bar{R} \cdot \overline{qKT}}} = s + \overline{qJT} + \overline{q\bar{R}(\bar{q} + \bar{K} + \bar{T})} = \\ &= s + \overline{qJT} + \overline{q\bar{R}(\bar{K} + \bar{T})}. \end{aligned}$$

On peut remarquer que la variable de sortie Q est exprimée par l'intermédiaire de Y_1 et Y_3 en fonction des trois variables primaires : $J, \bar{K}$ et T .

Le pas suivant dans l'analyse du circuit c'est la détermination du „symbole de marque“ [4] qui comprend, tout comme le tableau de vérité, toute information concernant les propriétés du circuit analysé, sous forme concentrée.

L'index supérieur de ce symbole représente la file des variables d'entrée, en commençant avec les secondaires et puis les primaires, séparés en groupes de trois variables, chacun pour faciliter la conversion binaire—octale des combinaisons des valeurs binaires de ces variables.

L'index inférieur du symbole comprend le fil des numéros octales, qui expriment des valeurs binaires des variables d'entrée pour lesquelles la variable de sortie est „1“ logique. Ces numéros, qui expriment des états possibles du circuit, s'appellent „numéros d'état“.

Dans notre exemple nous précisons l'index supérieur par le fil : $qSR : JKT$.

Pour trouver les numéros d'état qui correspondent aux termes de la forme canonique disjonctive de l'expression logique, on détermine tout d'abord les groupements de numéros d'état „adjacent“, qui correspondent à chaque terme de la forme normale disjonctive (1). Ces groupements correspondent aux „zones“ dans les diagrammes V e i t c h-K a r n a u g h.

Deux numéros d'état sont adjacents si, écrits sous forme binaire, elles diffèrent par un seul bit du même ordre. L'ordre représente la différence d'adjacence qui, en code octal, peut être : 1, 2, 4, 10, 20, 40, 100, ... etc.

Le premier numéro d'état d'un groupement adjacent est obtenu en transformant en code octal le numéro binaire résulté après le remplacement des variables non-niées par „1“ logique dans l'index supérieur et des variables niées ou inexistantes dans le terme analysé, par „0“ logique.

Le groupement adjacent est constitué par des numéros qui s'obtiennent en considérant les différences d'adjacences, dans leur ordre ascendant, différences qui correspondent aux variables qui manquent du terme analysé. Donc on écrit (tout comme on expose systématiquement ci-dessous) le numéro d'état adjacent au premier, (déterminé de la manière décrite ci-dessus), les deux numéros d'état suivants et adjacents aux premiers deux, puis les quatre numéros d'état suivants adjacents aux premiers quatre, etc.

Dans notre exemple, les résultats ont, du point de vue synthétique, l'aspect suivant :

On peut décrire de la même manière d'autres séquences de travail, commandées par exemple par S et/ou R .

Il est intéressant à mettre en évidence l'instabilité du circuit manifestée dans le cas où $J=K=T=1$, qui résulte de la séquence :

$$\frac{0}{0} \xrightarrow{J=K=1} \frac{1}{0} \xrightarrow{T=1} \left(\frac{7}{1}\right) \xrightarrow{q=1} \left(\frac{47}{0}\right) \xrightarrow{q=0} \left(\frac{7}{1}\right) \xrightarrow{q=1} \dots$$

Cette déficience de la mémoire est éliminée, tout comme dans l'exemple suivant, dans le cas du bistable master-slave.

3. Analyse du bistable CDB-472 (CDB-476)

La structure logique de ce circuit intégré roumain est présentée dans la Fig. 2.

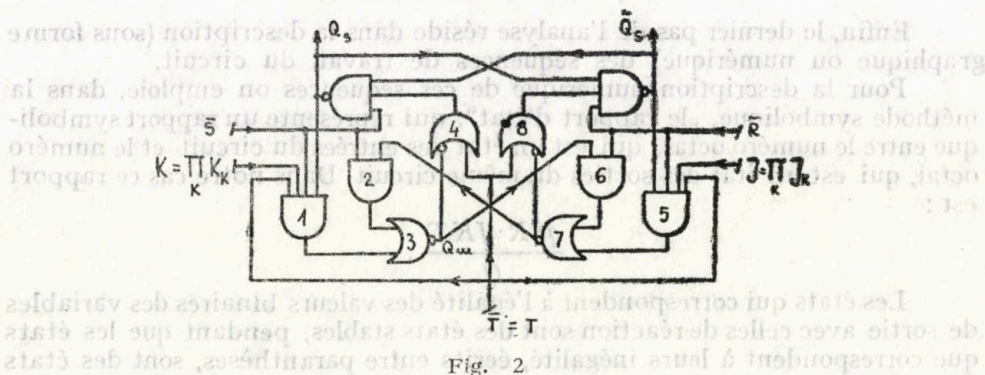


Fig. 2

Dans ce cas le circuit a deux variables de sortie : Q_s (slave) et Q_m (master), auxquelles on ajoute les variables de réaction, q_s et q_m . L'inspection du circuit logique permet d'écrire :

$$Q_m = Y_s = \bar{Y}_1 + \bar{Y}_2 = \bar{Y}_1 \bar{Y}_2; Y_1 = q_s \bar{S} K \bar{T}; Y_2 = \bar{S} Y_7; Y_7 = \bar{Y}_5 \bar{Y}_6; \\ Y_5 = \bar{q}_s \cdot J \bar{R} \bar{T}; Y_6 = q_m \bar{R},$$

ou finalement :

$$Q_m = S + \bar{T} J \bar{R} \bar{q}_s + q_m \bar{R} (T + \bar{K} + S + \bar{q}_s).$$

De la même manière, pour Q_s on obtient l'expression

$$Q_s = S + q_m (T + q_s \bar{R}) + q_s \bar{R} \bar{T}.$$

En formant les groupements de numéros d'état adjacents, tout comme dans l'exemple antérieur, il résulte le suivant symbole de marque du circuit :

$$Q_m, Q_s = \bigcup_{Q_m \cdot Q_s}^{q_m \cdot q_s} SR \cdot JKT$$

$$Q_m = 1: (4; 6; 20 \dots 37; 60 \dots 107; 120 \dots 141; 143 \dots 145;$$

$$147; 160 \dots 177)$$

$$Q_s = 1: (20 \dots 40; 42; 44; 46; 60 \dots 77; 101; 103; 105;$$

$$107; 111; 113; 115; 117; 120 \dots 147; 151; 153;$$

$$155; 157; 160 \dots 177)$$

Le rapport d'état utilisé dans la description symbolique des séquences de travail du circuit est :

$$\frac{q_m \cdot q_s SR \cdot JKT}{Q_m Q_s}$$

Les séquences principales de travail du circuit, dans le cas où $T' = \bar{T}$, sont :

a) Inscription normale :

$$\frac{1}{0} \xrightarrow{J=1} \frac{5}{6} \xrightarrow{T'=1} \left(\frac{4}{2} \right) \xrightarrow{q_m=1} \frac{104}{2} \xrightarrow{T'=0} \left(\frac{105}{3} \right) \xrightarrow{q_s=1} \frac{145}{3} \xrightarrow{J=0} \frac{141}{3}$$

On peut remarquer que pour $T'=1$ on obtient $Q_m=1$ et pour $T'=0$, on obtient $Q_s=1$.

b) Inscription forcée :

$$\frac{1}{0} \xrightarrow{S=1} \left(\frac{21}{3} \right) \xrightarrow{q_m=1} \frac{161}{3} \xrightarrow{T'=1} \frac{160}{3} \xrightarrow{S=0} \frac{140}{3} \xrightarrow{T'=0} \frac{141}{3}$$

On peut remarquer que pour $S=1$ on obtient $Q_m=Q_s=1$ sans tenir compte de la valeur du „clock“ T' .

c) La priorité de S par rapport à K est exprimée par la séquence :

$$\frac{141}{3} \xrightarrow{S=1} \frac{161}{3} \xrightarrow{K=1} \frac{163}{3} \xrightarrow{T'=1} \frac{162}{3}$$

d) L'effacement normal :

$$\frac{141}{3} \xrightarrow{K=1} \frac{143}{3} \xrightarrow{T'=1} \left(\frac{142}{1} \right) \xrightarrow{q_m=0} \frac{42}{1} \xrightarrow{T'=0} \left(\frac{43}{0} \right) \xrightarrow{q_s=0} \frac{3}{0} \xrightarrow{K=0} \frac{1}{0}$$

On peut remarquer que pour $T'=1$ on obtient $Q_m=0$ et pour $T'=0$ on a aussi $Q_s=0$.

e) L'effacement forcé :

$$\frac{141}{3} \xrightarrow{R=1} \left(\frac{151}{1} \right) \xrightarrow{q_m=0} \left(\frac{51}{0} \right) \xrightarrow{q_s=0} \frac{11}{0} \xrightarrow{T'=1} \frac{10}{0} \xrightarrow{R=0} \frac{0}{0} \xrightarrow{T'=0} \frac{1}{0}$$

Pour $R=1$ on obtient $Q_m=Q_s=0$, sans tenir compte de la valeur du „clock“ T' .

f) La priorité de R par rapport à J est exprimée par la séquence :

$$\frac{1}{0} \xrightarrow{R=1} \frac{11}{0} \xrightarrow{J=1} \frac{13}{0} \xrightarrow{T'=1} \frac{12}{0}$$

g) Le comportement du circuit en condition de simultanéité ($J=K=1$), est :

$$\begin{aligned} \frac{7}{0} \xrightarrow{T'=1} \left(\frac{6}{2}\right) \xrightarrow{q_m=1} \frac{106}{2} \xrightarrow{T'=0} \left(\frac{107}{3}\right) \xrightarrow{q_s=1} \frac{147}{3} \xrightarrow{T'=1} \left(\frac{146}{1}\right) \rightarrow \\ \xrightarrow{q_m=0} \frac{46}{1} \xrightarrow{T'=0} \left(\frac{47}{0}\right) \xrightarrow{q_s=0} \frac{7}{0} \rightarrow \dots \end{aligned}$$

On peut noter que pour $J=K=1$, le „clock“ T' se retrouve à la sortie du „slave“ et que le circuit ne garde plus l'instabilité spécifique au bistable JK , analysé ci-dessus.

4. Conclusions

1. L'analyse par la méthode symbolique permet la description en détail des séquences de travail d'un circuit séquentiel et la détection de ses éventuelles déficiences.

2. On peut analyser les circuits définis par un grand nombre de variables d'entrée et de sortie.

3. La méthode étant numérique, on peut établir un algorithme d'analyse programmable sur un ordinateur.

Reçue le 19 mai 1983

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ANALIZA PRIN METODA SIMBOLICĂ A UNOR CIRCUITE INTEGRATE DIGITALE

(Rezumat)

Se expune o metodă nouă de analiză a circuitelor secvențiale, aplicabilă și structurilor integrate, luând ca exemplu un circuit din familia CDB.

Metoda prezintă avantajul unei descrieri amănunțite a secvențelor de lucru a circuitului, este aplicabilă unor circuite cu număr mare de variabile de intrare și ieșire și se pretează programării pe calculator.

THE COMPENSATION OF THE COMPONENTS GIVEN BY THE SUBCARRIER THIRD HARMONIC IN STEREO CODERS

BY

GH. MAXIM, I. BOGDAN and V. FURNICĂ

1. Introduction

When using stereophonic broadcasting, the two sound sources signals, $L(t)$ and $R(t)$, are transmitted as a multiplex stereo signal (MSS) :

$$(1) \quad v(t) = L(t) + R(t) + [L(t) - R(t)] \sin 2\pi f_{sp}t + P \sin 2\pi(\frac{1}{2}f_{sp})t,$$

where : $f_{sp} = 38$ KHz — the subcarrier frequency ; $\frac{1}{2}f_{sp} = 19$ KHz — the pilot frequency ; P — the pilot amplitude [1].

The latest applications use analogue multiplexing to obtain the MSS from the primary audio signals, named coding operation [1]...[4]. The output of the analogue multiplexer (MUX) contains, in addition to the useful components as given by (1), other components as given by :

$$(2) \quad [L(t) - R(t)] \sin 2\pi(2n+1)f_{sp}t, \quad (n=1, 2, 3, \dots),$$

and a low-pass-filter (LPF) must be provided at the MUX output to eliminate them. This filter must carry out very severe requirements [5], namely : to introduce a less than 1% attenuation in the MSS frequency band (0...53 KHz), to have a linear phase—frequency characteristic and, in the same time, to produce a great attenuation (greater than 30 dB) of the unwanted components (mainly those given by the third harmonic of the subcarrier frequency).

This paper presents a method to compensate the components given by the subcarrier third harmonic, that are the most difficult to attenuate by the filter ; thus, the filter requirements are less hard and a very weak cross induction may be obtained during coding operation.

2. Analogue Multiplexing

The skeleton diagram of the used multiplexer is given in Fig. 1. The transistors T_2 - T_3 and T_5 - T_6 are matched (in an integrated circuit) and the current sources T_1 , T_4 are (for instance current) controlled in antiphase by a rectangular shaped signal with duty cycle 0.5 and 38 KHz frequency, generated by TTL circuits.

The two current sources currents are :

$$(3) \quad i_1(t) = I_0 y_1(t), \quad i_2(t) = I_0 y_2(t),$$

where I_0 is a constant current and the balance functions y_1 and y_2 are defined as following :

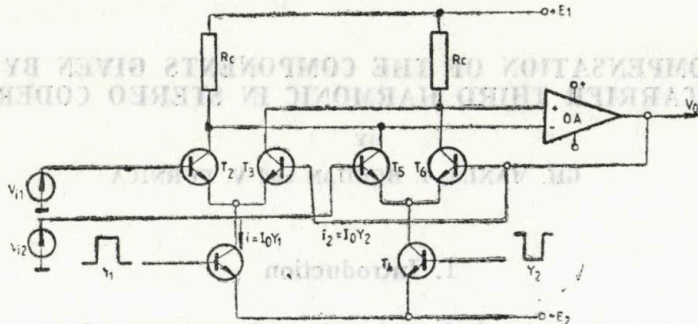


Fig. 1. — Multiplexer. Skeleton diagram.

$$(4) \quad \begin{cases} y_1 = 1 \text{ for } t \in [0, T/2] + nT \\ y_1 = 0 \text{ for } t \in [T/2, T] + nT, \\ y_2(t) = 1 - y_1(t), \quad (n = 1, 2, 3, \dots), \quad T = 1/(38 \text{ KHz}). \end{cases}$$

During the time intervals $[0, T/2] + nT$ the transistors pair T_5-T_6 is cut off and the transistors pair T_2-T_3 is fed through emitters with the constant current I_0 . The OA output is $v_0 = v_{i1}$.

During the time intervals $[T/2, T] + nT$ the transistors pair T_2-T_3 is cut off and the transistors pair T_5-T_6 is fed through emitters with the constant current I_0 . The OA output is $v_0 = v_{i2}$.

Thus, it may be written:

$$(5) \quad v_0(t) = v_{i1}(t)y_1(t) + v_{i2}(t)y_2(t).$$

Using Fourier series for $y_1(t)$ and $y_2(t)$, v_0 becomes :

$$(6) \quad \begin{aligned} v_0(t) = & \frac{1}{2} [v_{i1}(t) + v_{i2}(t)] + \frac{2}{\pi} [v_{i1}(t) - v_{i2}(t)] \sin 2\pi f_{sp} t + \\ & + \frac{2}{3\pi} [v_{i1}(t) - v_{i2}(t)] \sin 3(2\pi f_{sp})t + \\ & + \frac{2}{5\pi} [v_{i1}(t) - v_{i2}(t)] \sin 5(2\pi f_{sp})t + \dots \end{aligned}$$

It may be seen from (6) that the MUX output would contain the MSS components if $v_{i1}(t)$ and $v_{i2}(t)$ were $L(t)$ and $R(t)$, respectively (s. (1)); the pilot signal must be added afterwards. Besides useful components $v_0(t)$ contains other components too, round about the subcarrier odd harmonics. These ones must be eliminate by a LPF.

3. The Principle of Compensation of the Components given by the Subcarrier Third Harmonic

This principle is illustrated in Fig. 2. In addition to the main multiplexer MUX1 controlled as in Fig. 1, an auxiliary multiplexer MUX3 is used, this one controlled, also, by the v_{i1} and v_{i2} signals and utilising the balance functions y'_1 and y'_2 defined in the same manner as y_1 and y_2 , but having the frequency $3f_{sp}$.

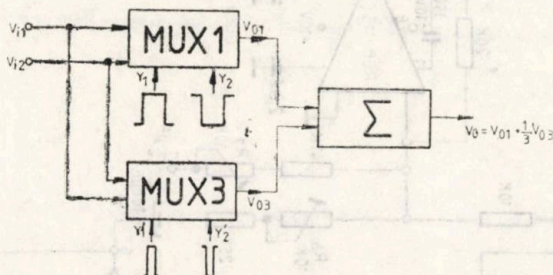


Fig. 2. — The principle of compensation of the components given by the subcarrier third harmonic.

The MUX3 output voltage is :

$$(7) \quad v_{o3}(t) = \frac{1}{2} [v_{i2}(t) + v_{i1}(t)] + \frac{2}{\pi} [v_{i2}(t) - v_{i1}(t)] \sin 2\pi(3f_{sp})t + \\ + \frac{2}{3\pi} [v_{i2}(t) - v_{i1}(t)] \sin 3.2\pi(3f_{sp})t + \dots$$

The sum of the two multiplexers output voltages, with appropriate balancing coefficients, is :

$$(8) \quad v_o(t) = v_{o1}(t) + \frac{1}{3} v_{o3}(t) = \frac{2}{3} [v_{i1}(t) + v_{i2}(t)] + \frac{2}{\pi} [v_{i1}(t) - v_{i2}(t)] \sin 2\pi f_{sp}t + \\ + \frac{2}{5\pi} [v_{i1}(t) - v_{i2}(t)] \sin 5(2\pi f_{sp})t + \dots$$

It may be noticed from (8) that the components given by the subcarrier third harmonic have been eliminated. As a result the LPF must eliminate the components given by the subcarrier harmonics greater than 5 and its design requirements become easier.

In order to obtain the useful components amplitudes ratio at the adder output as given by (1), the multiplexers input signals must be as following :

$$(9) \quad v_{i1}(t) = L(t) - pR(t), \quad v_{i2}(t) = R(t) - pL(t), \quad \text{with } p = \frac{\pi - 3}{\pi + 3}.$$

These signals may be obtained without circuit diagram complications during L and R individual processing.

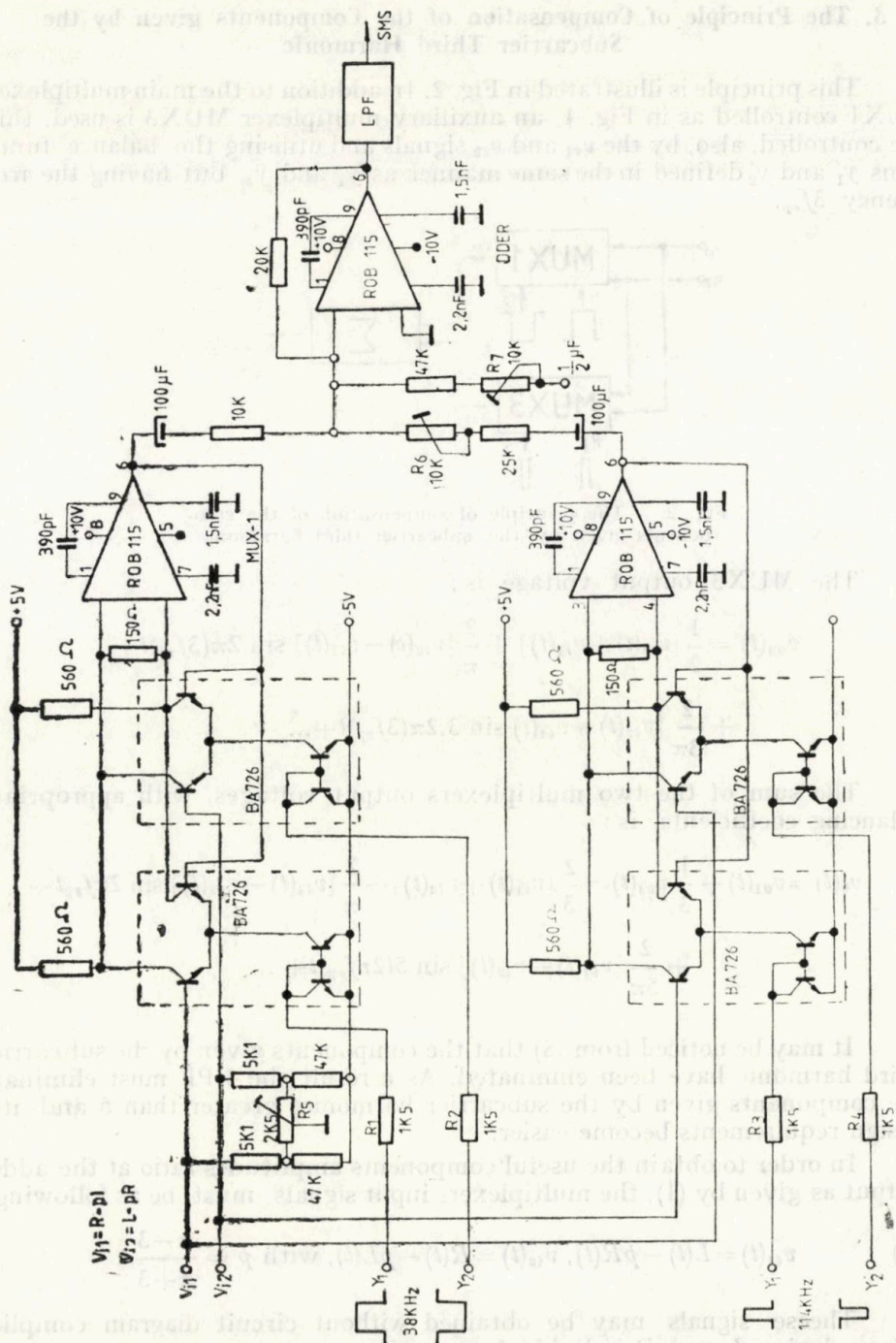


Fig. 3. — Stereo Coder with Compensation of the Subcarrier Third Harmonic. Circuit diagram.

If L and R signals were introduced directly to the multiplexers inputs, the cross induction attenuation would be limited at

$$(10) \quad D = 20 \lg \frac{\pi+3}{\pi-3} \approx 33 \text{ dB.}$$

4. Experimental Results

The coding principle presented before was experimentally studied by means of the circuit presented in Fig. 3.

The balance functions y and y' were obtained by division with 12 and 4, respectively, of a 456 KHz clock signal generated by a crystal resonator oscillator. These circuits have an usual diagram and are not represented in Fig. 3. Neither the circuits producing the multiplexers' input signals are represented in Fig. 3, because they are included in the preemphasis circuits.

The multiplexers are realised by means of $\beta A726$ integrated circuits. The balance signals are applied to the $\beta A726$ current sources by means of $R_1 \dots R_4$ bleeder resistor.

The integrated circuit ROB 115 is used as an OA in the multiplexers and the adder; this allowed to obtain a linear phase to frequency characteristic and a better than 1% uniformity of the attenuation to frequency characteristic in the whole 0...53 KHz frequency domain.

The pilot signal level at the input of the adder may be adjusted by means of the R_7 potentiometer.

A third order Cauer filter with a 100 KHz cut-off frequency is used as LPF.

It was noticed by means of a selective milivoltmeter that by a correct adjusting of the adder balance ratio (by means of R_6) the level of the components given at the adder output by the third harmonic may be maintained less than 45 dB for audio frequencies between 100 Hz and 10 KHz.

The cross induction was measured by means of a decoder and through the method described in [1], [5]. The L and R channels separation was better than 50 dB for audio frequencies between 100 Hz and 10 KHz.

The output subcarrier level may be maintained less than 50 dB by a correct adjusting of R_5 .

5. Conclusions

The disturbing components given by the subcarrier third harmonic in the analogue multiplexers from stereocoders may be compensated by means of an auxiliary multiplexer operating on the third harmonic, so that the output LPF requirements become easier and a better than 50 dB separation between L and R channels may be obtained with a quite simple circuit.

Received, August 1, 1983

Polytechnic Institute, Jassy,
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COMPENSAREA COMPONENTELOR DATE DE ARMONICA A TREIA A SUBPURTĂTOAREI ÎN CODOARELE STEREO

(Rezumat)

Se pune în evidență teoretic și experimental faptul că, în codoarele stereo cu multiplexare analogică, componentele parazite date de armonica a treia a subpurtătoarei pot fi compensate folosind un multiplexor suplimentar pe armonica a treia. Pe această cale se ușurează cerințele impuse filtrului de la ieșirea multiplex și se creează condiții pentru obținerea unei separări între canale mai bună de 50 dB, fără complicații deosebite de schemă.

5. Conclusions

The disturbing components given by the subcarrier third harmonic in the analogue multiplexers from stereocoders may be compensated by means of an auxiliary multiplexer operating on the third harmonic, so that the output LTP requirements become easier and a better than 50 dB separation between A and B channels may be obtained with a quite simple circuit.

A METHOD FOR THE DETERMINATION OF THE INPUT BIAS CURRENTS I_B^+ , I_B^- OF OPERATIONAL AMPLIFIERS WITH INPUT BIPOLAR TRANSISTORS

BY

C. ORITĂ, LAURENȚIU DIMITRIU, I. CLEJU and LUCIUS DUMITRESCU

1. Introduction

The input bias currents I_B^+ , I_B^- and input offset current $I_{\text{off}} = I_B^+ - I_B^-$ of the operational amplifiers are very interesting for users in many applications. Therefore the manufacturers include these in data sheets extreme values, obtained in direct and indirect measurements.

The measurement of these currents is carried out [1]...[4], by means of the voltages caused by I_B^+ , I_B^- currents across resistors with well-known values. Since these voltages are influenced by the offset voltage, V_{off} , this parameter must be determined in a previous measurement [1], [2] or there are used circuits where V_{off} is eliminated by means of two successive determinations [3], [4].

2. Generalities

For a normal bias of a npn planar transistor, the base current can be written as [5]:

$$(1) \quad I_B = I_{PE} + I_{VB} + I_{SB},$$

where: I_{PE} is the hole current, injected from base to emitter, I_{VB} — the volume recombination current, I_{SB} — the surface recombination current.

The three components of the base current have the following expressions:

$$(2) \quad I_{PE} = I_{PEs} \exp\left(q \frac{V_{BE}}{KT}\right), \quad (3) \quad I_{VB} = I_{VBs} \exp\left(\frac{V_{BE}}{KT}\right),$$

$$(4) \quad I_{SB} = I_{SBs} \exp\left(q \frac{V_{BE}}{KT}\right),$$

where I_{PES} , I_{VBS} , I_{SBs} are constant for a certain temperature. We can write:

$$(5) \quad I_B = I_{BS} \exp \left(q \frac{V_{BE}}{KT} \right),$$

where $I_{BS} = \text{const.}$ for $T = \text{const.}$

For small deviations v_{BE} around the bias voltage V_{BE} and using (5) it results:

$$(6) \quad \frac{|\Delta i_B|}{I_B} \approx \frac{q |\Delta V_{BE}|}{KT}.$$

The smaller deviation Δv_{BE} in relation to bias voltage V_{BE} , the more exactly expression (6) is.

For the transistors of the input stage of an operational amplifier with a very large open loop gain, the expression (6) can be considered an equality. The expression (6) is obtained using Taylor series expansion for the expression (5) and in this situation the other terms of the series are very small than the first two. If we suppose a perfectly balanced input stage of an operational amplifier with input bipolar transistors we can write:

$$(7) \quad |\Delta v_o| = 2a_o |\Delta v_{BE}|,$$

where a_o is the midband open loop gain and $|\Delta v_o|$ is the output voltage deviation due to the differential input voltage, $|\Delta v_{id}| = 2|\Delta v_{BE}|$.

Using the expressions (6), (7), we can write:

$$(8) \quad \left| \frac{\Delta i_B}{I_B} \right| = \frac{|\Delta v_o|}{2a_o} \cdot \frac{q}{KT}.$$

In the case of an operational amplifier with $a_o \geq 10\,000$, working at room temperature and for variations of output voltage $|\Delta v_o| \leq 1\text{ V}$, it results:

$$(9) \quad \left| \frac{\Delta I_B}{I_B} \right| \% \leq 0,2\%.$$

The last inequality point out that an operational amplifier with input bipolar transistors and $a_o \geq 10\,000$ has insignificant variations of input currents i_B^+ , i_B^- in relation to input bias currents I_B^+ , I_B^- for significant variations of output voltage. We can consider that the input bias currents has constant values for significant variations of output voltage. Based on this conclusion, we propose a method for the determination of the I_B^+ , I_B^- input bias currents. This method will be justified by means of the circuit diagram from Fig. 1 and diagrams from Fig. 2.

3. Principle of Method

The capacitors C^+ , C^- from the Fig. 1 together with the switches a , b , c , determine the precision of this method by their leakage currents. For a good accuracy, these leakage currents must be much smaller than the

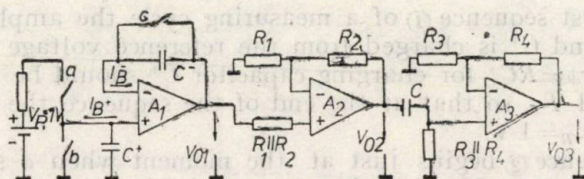


Fig. 1

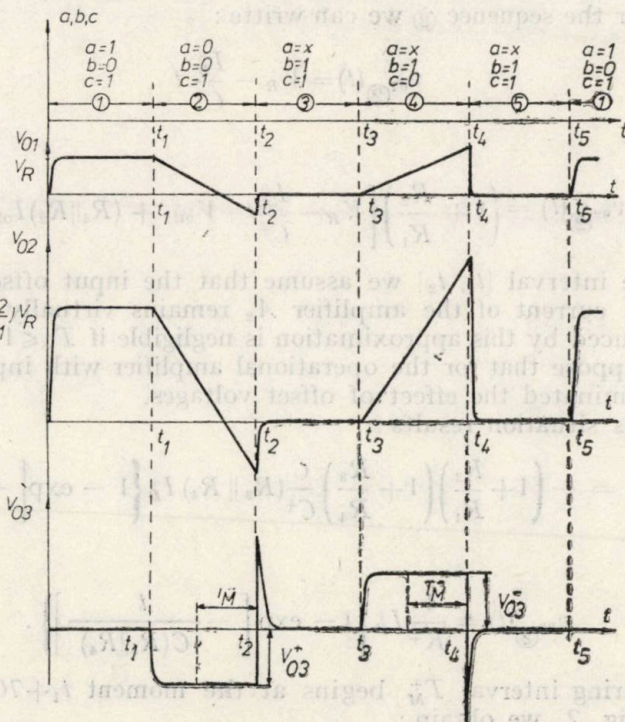


Fig. 2

input bias currents of the tested operational amplifier ($I_1 \leq 100 I_B^+$, $I_1 \leq 100 I_B^-$). Using teflon dielectric capacitors for C^+ , C^- and MOSFET's for switches a , b , c , whose leakage currents are smaller than 0.1 nA in the OFF state at room temperature, the method becomes useful for operational amplifiers with input bias currents greater than 10 nA.

The principle of the method is presented supposing that C^+ , C^- , a , b , c , are ideal in Fig. 1. On the other hand we assign the value „1“ for a switch in the ON state, „0“ for a switch in the OFF state and „X“ for random state.

The three switches of the circuit are controlled by an external driving system.

The states of the three switches are shown in the sequences ①...⑤ (Fig. 2). The duration T_s of every sequence is much greater than the rise time for the step response of the tested amplifier A_1 .

In the first sequence ① of a measuring cycle the amplifier A_1 works as a follower and C^+ is charged from the reference voltage $V_R=1$ V. The time constant $\tau_+ = RC^+$ for charging capacitor C^+ should be much smaller than the period T_s , so that at the end of the sequence the voltage across C^+ should be $V_R=1$ V.

The sequence ② begins just at the moment when a switches OFF. A_1 remains a follower and v_{01} tends to negative values due to the discharging of C^+ by I_B^+ . Since for variations of v_{01} around the origin, I_B^+ remains virtually constant, for the sequence ② we can write:

$$(10) \quad v_{01}^{(2)}(t) = V_R - \frac{I_B^+}{C^+} t$$

and

$$(11) \quad v_{02}^{(2)}(t) = \left(1 + \frac{R_2}{R_1}\right) \left[V_R - \frac{I_B^+}{C^+} t + V_{off2} + (R_1 \parallel R_2) I_{off2} \right].$$

In the interval $[t_1, t_2]$ we assume that the input offset voltage and input offset current of the amplifier A_2 remains virtually constant. The error introduced by this approximation is negligible if $T_s \leq 1$ s.

We suppose that for the operational amplifier with input FET's, A_3 , we have eliminated the effect of offset voltages.

In this situation results:

$$(12) \quad v_{03}^{(2)}(t) = - \left(1 + \frac{R_2}{R_1}\right) \left(1 + \frac{R_3}{R_4}\right) \frac{C}{C^+} (R_3 \parallel R_4) I_B^+ \left\{ 1 - \exp \left[- \frac{t}{C(R_3 \parallel R_4)} \right] \right\},$$

or

$$(13) \quad v_{03}^{(2)}(t) = \frac{1}{K^+} I_B^+ \left\{ 1 - \exp \left[- \frac{t}{C(R_3 \parallel R_4)} \right] \right\}.$$

If the measuring interval T_M^+ begins at the moment $t_1 + 7C(R_1 \parallel R_2)$ as is shown in Fig. 2, we obtain:

$$1 - \exp \left[- \frac{t}{C(R_3 \parallel R_4)} \right] \geq 1 - 9.12 \times 10^{-4}.$$

For T_M^+ interval we can write

$$(14) \quad I_B^+ = K^+ V_{03}^+,$$

where $K^+ = \text{co. st.} < 0$.

The resistors R_1 , R_4 and the capacitors C^+ , C^- can be chosen so that in the expression (14) $|K^+| = 10^{-9}$ and $|V_{03}^+| = 1 \dots 10$ V for measuring currents $i = 10 \dots 100$ nA.

To extend at 100 nA... $1 \mu\text{A}$ the measuring range, it must increase $|K^+|$ ten times by reducing also ten times the capacitor C^+ .

During T_M^+ interval, the value of dc voltage V_{03}^+ can be digitally measured. So we can obtain on the display directly the value of I_B^+ current.

In the sequence ③ the switch b switches ON and until the moment t_{3t} , v_{03} reach the value 0 V, so $v_{03}(t_3)=0$. After the moment t_3 begins the sequence ④. In this sequence c switches OFF and C^- integrates the current I_B^- . In the same way as at the sequence ②, we can write, if we consider $C^+=C^-$ and $T_M^+=T_M^-$:

$$(15) \quad I_B^- = K^- V_{03}^-$$

where $|K^+| = |K^-|$.

The value of I_B^- current can be also displayed on a digital display.

4. Conclusions

The proposed method has the advantage of simplicity and no taking into consideration the offset voltage V_{off1} and V_{off2} of tested amplifier, A_1 , and the amplifier A_2 .

It may be used for determining the input bias currents I_B^+ and I_B^- of the operational amplifiers with input bipolar transistors and the midband open loop gain $a_0 \geq 10^4$.

Received, July 4, 1983

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METODĂ PENTRU DETERMINAREA CURENȚILOR I_B^+ , I_B^- AI AMPLIFICATOARELOR OPERAȚIONALE CU TRANZISTOARE BIPOLARE LA INTRARE

(Rezumat)

Se propune o metodă pentru determinarea curenților de intrare I_B^+ , I_B^- ai amplificatoarelor operaționale cu tranzistoare bipolare la intrare, bazată pe proprietățile specifice ale amplificatoarelor operaționale. Metoda nu ia în considerare decalajul de tensiune V_{off} de la intrare și se remarcă prin simplitate. Se folosește pentru determinarea curenților de intrare la amplificatoarele operaționale cu $I_B \geq 10$ nA și $a_0 \geq 10^4$.

In the sequence of the switch S_2 and until the moment t_2 the value of V_{A_2} is zero. After the moment t_2 begins the register C_2 in this sequence a switch S_2 and C_2 integrates the current I_{A_2} . In the same way as at the sequence S_1 we can write: if we consider $C_2 = 1$ and $V_{A_2} = V_{A_2}$.

$$I_{A_2} = K_2 \cdot T_{A_2} \quad (13)$$

where $K_2 = A_2$. The value of I_{A_2} current can be also displayed on a digital display.

4. Conclusions

The proposed method has the advantage of simplicity and no taking into consideration the offset voltage V_{off} and T_{off} of tested amplifier A_2 and the amplifier A_1 . It may be used for determining the input bias currents I_{B_1} and I_{B_2} of the operational amplifiers with input bipolar transistors and the midband open loop gain $A_0 > 10^4$.

Received July 4, 1981
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METODA PENTRU DETERMINAREA CURENTILOR I_{B_1} SI I_{B_2} SI AMPLIFICATORILOR OPERATIONALI CU TRANSISTORI BIPOLARI LA INTRARE

(Rezumat)

Se prezinta o metoda pentru determinarea curentilor de intrare I_{B_1} si I_{B_2} ai amplificatorilor operationali cu tranzistori bipolari la intrare, bazata pe o proprietate specifica ale amplificatorilor operationali. Aceasta nu este necesara descalibrarea tensiunii V_{off} de la intrare si se realizeaza prin simplitate. Se indica si pentru determinarea conversiei de intrare la amplificare operationala cu $A_0 > 10^4$ si $A_0 > 10^4$.

C.D. 621.314.3

CIRCUITS DE COMMANDE POUR LE VARIATEUR TRIPHASÉ DE TENSION ALTERNATIVE À TROIS THYRISTORS

PAR

AL. HRITCU et M. LUCANU

1. Introduction

Le circuit de puissance du variateur triphasé de tension alternative à trois thyristors est représenté dans la fig. 1 [1]. Ce variateur est utilisé sur une charge triphasée symétrique qui a le point neutre accessible. L'utilisation des trois thyristors dans le circuit de puissance détermine une augmentation de la sollicitation des thyristors. La plus grande sollicitation d'un thyristor (qui pendant le fonctionnement du variateur est à commande continue) est de 50% environ plus grande que celui d'un thyristor dans un contacteur à six thyristors [2].

Le fonctionnement du variateur sur charge inductive et résistive est analysé dans [3], [4] où sont données les relations pour le calcul du circuit de puissance.

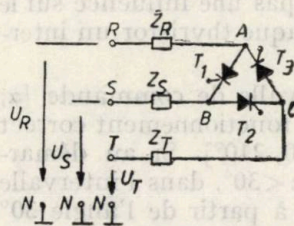


Fig. 1

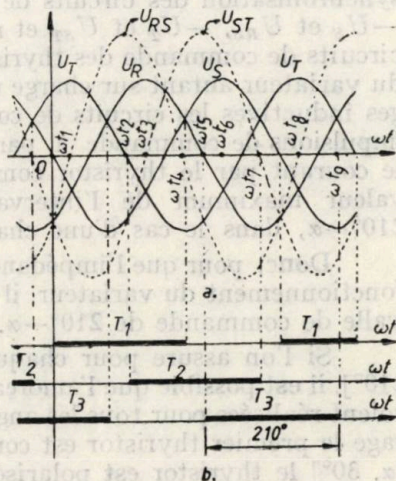


Fig. 2

Dans ce qui suit on analyse les conditions qu'on doit imposer aux circuits de commande des thyristors du variateur, en se proposant deux circuits de commande.

2. Les prescriptions imposées aux circuits de commande des thyristors

L'analyse des conditions qu'on doit imposer aux circuits de commande est effectuée dans le cas des formes d'onde des tensions du réseau représentées dans la Fig. 2 a. À l'état d'attente les thyristors $T1$, $T2$, $T3$ du variateur sont sollicités aux tensions U_{RS} , U_{ST} et respectivement U_{TR} . Pour l'amorçage du variateur il faut que le premier thyristor soit commandé d'une certaine façon, à savoir: $T1$ dans l'intervalle $[\omega t_1, \omega t_4]$, quand $U_{RS} > 0$, $T2$ dans l'intervalle $[\omega t_3, \omega t_7]$ quand $U_{ST} > 0$ et $T3$ dans l'intervalle $[\omega t_6, \omega t_9]$ quand $U_{TR} > 0$ (Fig. 2a).

Après l'amorçage du variateur l'intervalle de variation de l'angle de commande pour chaque thyristor doit assurer le réglage de la tension dans le domaine $0 \dots U_{V_{eff}}$, où U représente la valeur effective de la tension d'alimentation du variateur pour une phase du réseau triphasé.

En analysant le circuit de puissance du variateur on détermine que la valeur maximum de la tension est obtenue par la commande des thyristors $T1$, $T2$, $T3$ aux moments 0 , t_2 et respectivement t_5 , mais la valeur minimum est obtenue par la commande des mêmes thyristors aux moments t_4 , t_7 et respectivement t_9 .

Donc, les intervalles de variation qu'on doit assurer pour les angles de commande des thyristors $T1$, $T2$, $T3$ sont respectivement $[0, \omega t_4]$, $[\omega t_2, \omega t_7]$ et $[\omega t_5, \omega t_9]$. On définit l'angle de commande prescrit pour $T1$ à partir de 0 , pour $T2$ à partir de ωt_2 et pour $T3$ à partir de ωt_5 .

Les intervalles de variation des angles de commande, d'une grandeur de 210° , étant déterminés, il résulte que les tensions utilisées pour la synchronisation des circuits de commande des thyristors $T1$, $T2$, $T3$ sont $-U_S$ et U_{RS} , $-U_T$ et U_{ST} et respectivement $-U_R$ et U_{TR} . Il faut que les circuits de commande des thyristors déterminent un bon fonctionnement du variateur autant sur charge résistive que sur charge inductive. Aux charges inductives les circuits de commande doivent assurer aux thyristors des impulsions de commande à partir de l'angle prescrit jusqu'à l'angle duquel le courant par le thyristor commandé dépasse le courant de maintien. La valeur maximum de l'intervalle de commande pour des thyristors est $210^\circ - \alpha$, dans le cas d'une charge inductive.

Donc, pour que l'impédance de charge n'exerce pas une influence sur le fonctionnement du variateur, il faut assurer pour chaque thyristor un intervalle de commande de $210^\circ - \alpha$.

Si l'on assure pour chaque thyristor un intervalle de commande $[\alpha, 210^\circ]$, il est possible que l'amorçage du variateur et le fonctionnement correct soient réalisées pour tous les angles de l'intervalle $[0, 210^\circ]$. Si au démarrage le premier thyristor est commandé d'un angle $\alpha < 30^\circ$, dans l'intervalle $[\alpha, 30^\circ]$ le thyristor est polarisé inversement, mais à partir de l'angle 30° le thyristor commence à conduire, la commande étant effectuée dans l'intervalle $[\alpha, 210^\circ]$. Le thyristor suivant, aussi commandé à l'angle α , commence à conduire à partir du même angle α , en étant précédé par un thyristor en conduction, de cette façon les thyristors ne sont pas polarisés par des tensions de ligne du réseau.

3. La description des circuits de commande proposés

Les circuits de commande des thyristors ont été réalisés selon les schémas représentés dans la Fig. 3, où : *CS* — circuit de synchronisation, *GTLV* — générateur de tension linéaire variable, *CC* — circuit comparateur, *OF* — oscillateur final.

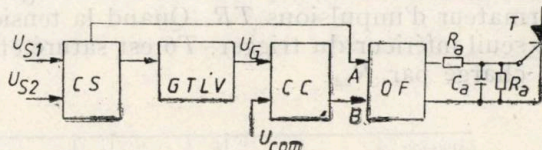


Fig. 3

Le circuit de commande réalisé selon ce schéma doit fonctionner pendant qu'au moins l'une des tensions de synchronisation U_{S1} , U_{S2} est positive, dans cette situation *CS* déclenchant le fonctionnement du circuit *GTLV*.

Quand la tension de sortie du circuit *GTLV* est $U_G < U_{com}$, le circuit *CC* établit à l'entrée *B* de l'*OF* une tension qui bloque le fonctionnement de cet étage. Quand $U_G > U_{com}$ le circuit *CC* établit à l'entrée *B* une tension qui permet le fonctionnement de l'*OF*, le thyristor étant commandé avec des impulsions de courte durée dans l'intervalle $[\alpha, 210^\circ]$.

Pendant que les deux tensions de synchronisation sont négatives, le circuit *CS* établit à l'entrée *A* de l'*OF* une tension qui arrête le fonctionnement de cet étage, indépendant de la tension établie à l'entrée *B*.

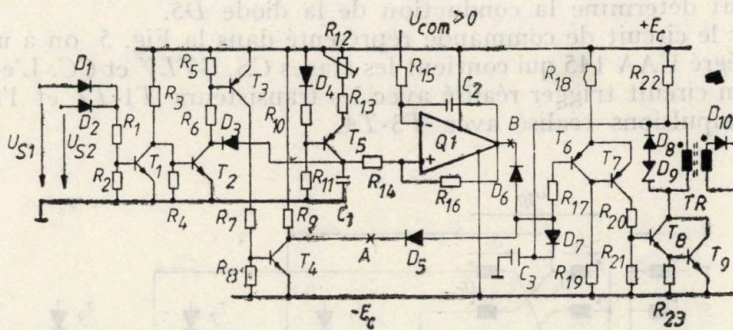


Fig. 4

Les étages du schéma de la Fig. 3 sont réalisés dans le circuit électronique représenté dans la Fig. 4, à savoir : *CS* est réalisé avec *T1-T2-T3-T4*, *GTLV*, avec *T5-C1*, *CC*, avec l'amplificateur *Q1* et *OF*, avec *T6-T7-T8-T9*. Pendant qu'au moins l'une des tensions U_{S1} , U_{S2} est positive, *T1* conduit en saturation et *T2-T3-T4* sont bloqués. L'état des transistors *T4*, *T2* détermine la polarisation inverse des diodes *D3* et *D5* et le condensateur *C1* est chargé avec un courant constant. Lorsque $U_{C1} < U_{com}$ la tension

à l'entrée B est $U_B = -E_c$, ce qui détermine la conduction de la diode D_6 . Ainsi le circuit trigger réalisé avec T_6 - T_7 est dans l'état T_6 saturé et T_7 bloqué, ce qui détermine le blocage des transistors T_8 T_9 et le chargement du condensateur C_1 vers la tension $+E_c$. Pendant que $U_{C1} > U_{om}$ la tension à l'entrée B devient $U_B = E_c$, la diode D_6 étant bloquée. Maintenant les thyristors du trigger sont dans l'état T_6 bloqué et T_7 saturé ce qui détermine la conduction des transistors T_8 et T_9 et la décharge du condensateur C_3 par le transformateur d'impulsions TR . Quand la tension du condensateur C_3 atteint le seuil inférieur du trigger, T_6 est saturé et T_7 bloqué et le condensateur est chargé par R_{22} .

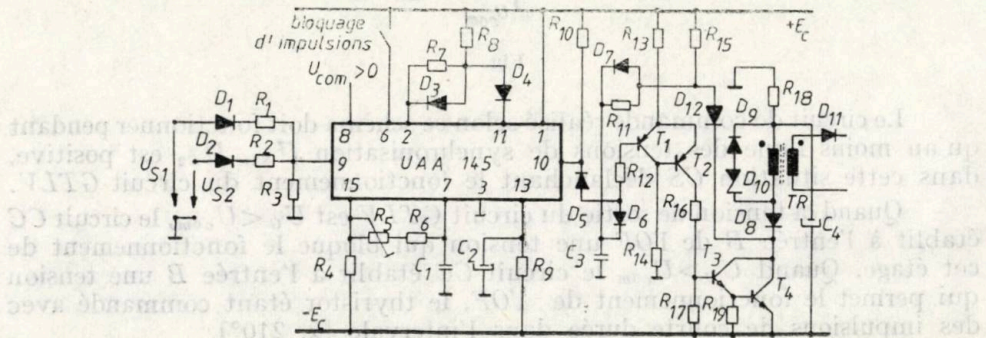


Fig. 5

Quand les tensions de synchronisation U_{S1} , U_{S2} deviennent toutes négatives, l'OF ne fonctionne pas parce que à l'entrée A la tension est $U_A = -E_c$, qui détermine la conduction de la diode D_5 .

Pour le circuit de commande représenté dans la Fig. 5 on a utilisé le circuit intégré UAA 145 qui contient les étages CS, GTLV et CC. L'étage OF contient un circuit trigger réalisé avec les transistors T_1 - T_2 et l'amplificateur d'impulsions réalisé avec T_3 - T_4 .

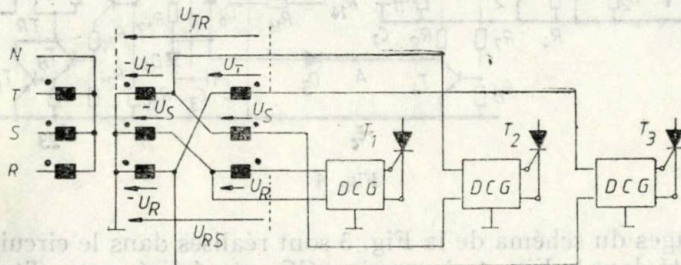


Fig. 6

Le fonctionnement des circuits OF représentés dans les fig. 4 et 5 est similaire avec la seule différence que dans la fig. 5 l'impulsion de commande n'est pas obtenue par le déchargement du condensateur C_3 par le transfor-

mateur d'impulsions, mais l'alimentation du transformateur à la tension $-E_c$, pendant le déchargement du condensateur C3 par R_{11} - R_{12} -T3 T4.

Les éléments R_7 - R_8 -D3- R_4 assurent le fonctionnement du circuit OF dans l'intervalle $[\alpha, 210^\circ]$; dans cet intervalle le circuit intégré UAA 145 détermine le blocage de la diode D5. Par l'alimentation du transformateur d'impulsions à la tension $-E_c$ et par C4 R_{18} on assure l'amorçage du circuit OF.

Donc, la synchronisation constitue le problème spécifique pour le circuit de commande du variateur à trois thyristors, les tensions nécessaires étant $-U_R$, $-U_S$, $-U_T$, U_{RS} , U_{ST} , U_{TR} . La synchronisation peut être réalisée avec un seul transformateur triphasé qui a six enroulements de sortie identiques. La connexion des enroulements du circuit de commande est montrée dans la Fig. 6, où DCG représente le circuit de commande d'un thyristor.

4. Conclusions

En tenant compte des particularités du variateur triphasé réalisé avec trois thyristors on a établi l'intervalle de variation de l'angle de commande des thyristors et le mode de synchronisation des circuits de commande.

On a proposé tant un circuit de commande avec les étages CS, CC et GTLV séparément réalisés, qu'un circuit de commande avec le circuit intégré UAA145 dont des éléments périphérique ont été modifiés pour satisfaire les prescriptions de commande.

Les schémas proposés se caractérisent par simplicité et par un fonctionnement sûr, le variateur étant convenable de ce point de vue aussi.

*Institut Polytechnique, Jassy,
Chaire d'Électronique et Calculateurs*

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CIRCUITE DE COMANDĂ PENTRU VARIATORUL TRIFAZAT DE TENSIUNE ALTERNATIVĂ CU TREI TIRISTOARE

(Rezumat)

Se analizează condițiile pe care trebuie să le îndeplinească dispozitivele de comandă a tiristoarelor unui variator trifazat de tensiune alternativă cu trei tiristoare, propunându-se în final două circuite de comandă ce îndeplinesc condițiile impuse, un dispozitiv utilizând circuitul integrat, specializat pentru comanda tiristoarelor, UAA 145.

POSSIBILITÉS D'ESSAI DES MOTEURS LINÉAIRES DE PUISSANCE MOYENNE

PAR

GH. LIVINȚ, LUCIAN CIOBANU et LORIN CANTEMIR

1. Introduction

L'essai des moteurs linéaires d'induction a pour but la détermination des caractéristiques mécaniques $F_t = f_1(v)$, des caractéristiques énergétiques, du rendement $\eta = f_2(v)$, du facteur de puissance $\cos \varphi = f_3(v)$, des courants, en vue d'optimiser la construction, d'améliorer les performances et de les tester dans des conditions variées de fonctionnement.

À présent on connaît plusieurs variantes de construction des bancs d'essai [1]... [3].

Pour des petites puissances on utilise des induits du moteur linéaire d'induction (MLI) sous forme de bande fixe ou mobile déplacée sur deux tambours [4], [5].

Pour des moyennes et grandes puissances l'induit du MLI a la forme d'un disque, entraîné d'habitude dans un mouvement de rotation par l'interaction électromagnétique entre celui-ci et l'inducteur du moteur linéaire [2], [3].

Dans [1] on présente une variante de construction où l'induit sous forme de disque est fixe et les inducteurs, placés sur une poutre, exécutent un mouvement de rotation suivant l'axe du disque induit sur la base de la même interaction électromagnétique.

Dans les bancs présentés ci-dessus la charge du MLI peut être modifiée, en général, à l'aide d'une machine de courant continu en modifiant le courant d'excitation et respectivement celui rotorique, lors de son fonctionnement freiné ou en régime de générateur.

La force développée par le MLI est déterminée dans ce cas sur la base de la puissance fournie par le générateur de courant continu.

2. La construction du banc d'essai

Le banc d'essai des moteurs linéaires de puissance moyenne réalisé (Fig. 1), est constitué d'un moteur linéaire d'induction aux inducteurs 1a et 1b fixés sur un cadre 2 et de l'induit en forme de couronne circulaire 3 qui exécute un mouvement de rotation autour de l'arbre 4, étant rigidisé par celui-ci par l'intermédiaire d'une plaque 5. Les inducteurs MLI peuvent être déplacés dans la direction du rayon $y-y'$ de l'induit 3 par l'intermédiaire du cadre 2 et du dispositif de guidage et de fixation 6; ils peuvent aussi exécuter un mouvement de rotation autour d'un axe vertical imaginaire $Z-Z'$ qui passe par le centre de symétrie des inducteurs, situé à l'intersection des diagonales des surfaces rectangulaires actives.

Pour charger le MLI on utilise une ou plusieurs couronnes circulaires 7, rigidisées par la plaque 5 à l'aide des barres 8. Sur la couronne 7 on applique une force de pression à l'aide des sabots de frein 9, fixés sur les poteaux

verticaux 10. On peut régler la force de pression pour chaque sabot par la compression ou le relachement d'un ressort calibré 11 muni d'un système qui indique la force de pression. On peut régler la force de freinage totale par le nombre des sabots qui pressent sur la couronne circulaire 7 ou, éventuellement, par le nombre de couronnes circulaires.

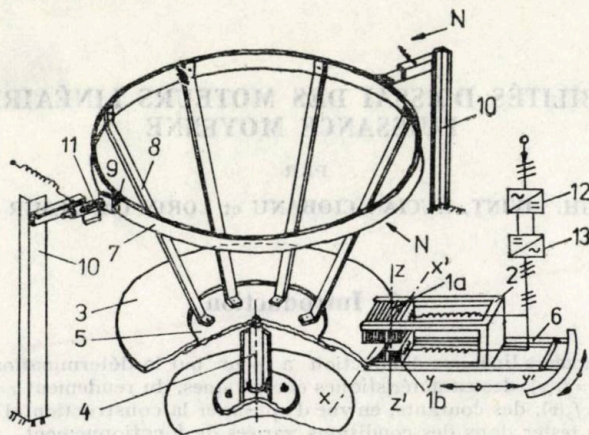


Fig. 1

Pour l'alimentation du MLI on utilise des sources réglables, ayant la possibilité de faire varier l'amplitude et la fréquence de la tension d'alimentation, constituée par le redresseur 12 et l'onduleur 13. La source d'alimentation réglable permet la modification de la fréquence entre 10 et 100 Hz et de la tension d'alimentation entre 20 et 400 V.

3. Détermination de l'effort résistant du MLI

L'effort résistant du MLI testé sur le banc est formé de deux composants: l'effort de freinage, F_f et la force de frottement avec l'air et dans les paliers, F_{a1} .

On peut calculer l'effort de freinage à l'aide de la relation :

$$(1) \quad F_f = \mu \sum_{i=1}^n N_i,$$

où : N_i — la force de pression sur le sabot i ; n — le nombre de sabots nécessaires pour charger le MLI ; μ — le coefficient de frottement entre les sabots et la couronne circulaire, que l'on peut calculer à l'aide de la relation :

$$(2) \quad \mu = \frac{17}{\sqrt{(0,1N + 0,25)(3,6v + 40)}},$$

avec : N — la force de pression normale, pour les valeurs moindres que 5 kN ; v — la vitesse tangentielle entre le sabot et la couronne 7, [km/h].

Le couple résistant à l'axe de rotation 4 devient :

$$(3) \quad M_{rJ} = \frac{1}{2} F_J D,$$

où D est le diamètre de la couronne circulaire 7; on peut adopter $D=6$ m.

Le diamètre de la couronne circulaire ayant une valeur importante il permet d'obtenir des couples résistants grands avec des forces de pression réduites.

Le couple déployé par le MLI dépend de la composante tangentielle, F , de la force déployée et du rayon R_v mesuré à partir du centre de symétrie des inducteurs jusqu'à l'axe de rotation :

$$(4) \quad M = F R_v$$

L'équation de l'équilibre mécanique peut être écrite soit par rapport à l'axe de rotation 4, soit par rapport à l'axe de symétrie longitudinale $x-x'$ du MLI. Donc :

$$(5) \quad M = M_r + J \left(\frac{d\omega}{dt} \right), \quad F = F_r + m \left(\frac{dv}{dt} \right),$$

où : J est le moment d'inertie de tous les éléments entraînés dans le mouvement de rotation par rapport à l'axe 4 ; v — la vitesse tangentielle au niveau du centre de symétrie des inducteurs ; ω — la vitesse angulaire ; F_r — la force résistante rapportée à l'axe $x-x'$. Dans des conditions d'équilibre ($v = \text{const.}$) on a $F = F_r$. Il en résulte la possibilité de déterminer la caractéristique mécanique $F = f(v)$ du MLI en prenant en considération toutes les composantes de la force F_r .

La force F_r est donnée aussi par la force de frottement des paliers et, respectivement, par la force de frottement contre l'air et par l'effort de freinage des sabots rapporté à l'axe $x-x'$:

$$(6) \quad F_r = F'_f + F_{al},$$

La force de frottement contre l'air, F_{al} , peut être déterminée par la méthode du lancement et l'on peut calculer d'une façon approximative à l'aide de la relation :

$$(7) \quad F_{al} = m_{\text{ech}}(a + bv + cv^2),$$

où : m_{ech} est la masse équivalente du mouvement de translation rapportée à l'axe $x-x'$; v — la vitesse tangentielle ; a, b, c — des constantes qui peuvent être déterminées par voie expérimentale.

L'effort de freinage des sabots rapporté à l'axe $x-x'$, F'_f , est calculé par la relation :

$$(8) \quad F'_f = \frac{F_J D}{D_y}.$$

La force déployée par le moteur linéaire conformément à (6) devient donc :

$$(9) \quad F = F_r = m_{ech}(a + bv + cv^2) + \mu \left(\frac{D}{D_y} \right) \sum_{i=1}^n N_i.$$

Compte tenu du fait que l'induit a la forme d'un disque en rotation et l'inducteur fixe du MLI a la forme de parallépipède ayant l'entrefer plan, il n'y a donc pas de concordance entre la vitesse du champ inducteur et les vitesses des différents points de l'induit. Certaines corrections sont nécessaires dans le calcul de la force déployée par le MLI.

4. Détermination des coefficients de correction

Les particularités de fonctionnement du MLI à inducteur plan-linéaire et à induit en disque à mouvement circulaire sont les suivantes :

a) la variation de la vitesse tangentielle des points du disque en fonction du rayon ;

b) la modification de la direction des vitesses tangentielles des points du disque par rapport à la vitesse v_1 de synchronisme du champ magnétique inducteur.

Ainsi, pour deux points de l'induit, qui correspondent à des rayons R_1 et R_2 , le rapport des vitesses tangentielles est égal au rapport des rayons.

Si on considère un point quelconque P sur l'induit disque (Fig. 2), entre la vitesse tangentielle réelle v_{Pt} et la vitesse v_1 du champ inducteur il y aura un angle θ qui se modifie avec le déplacement de l'induit. Pour un angle θ constant, la vitesse tangentielle synchrone v_{PT} est

$$v_{PT} = \frac{v_1}{\cos \theta}.$$

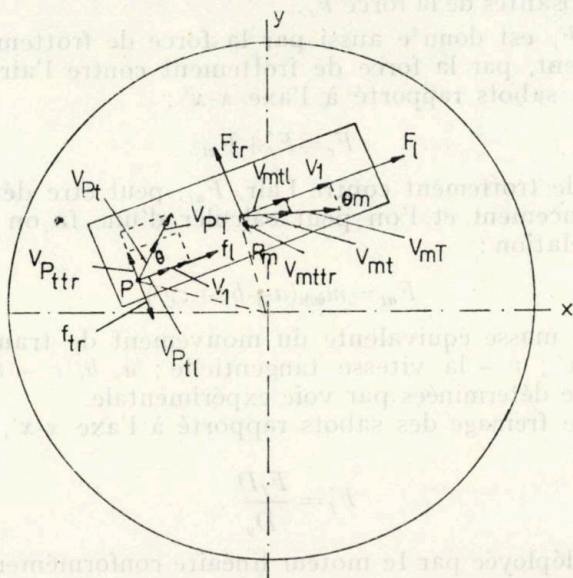


Fig. 2

Cette vitesse (v_{Pt}) peut être décomposée en deux composants : l'un selon la direction longitudinale v_{Ptl} , colinéaire avec la vitesse synchrone v_1 du champ inducteur et l'autre transversal, v_{Ptr} , perpendiculaire sur v_1 . Aux deux composants de la vitesse v_{Pt} correspondent les deux composants de la force déployée par le MLI dans le point P , notamment f_1 et f_{tr} .

En correspondance avec les dimensions finies des inducteurs MLI dans (7) on définit une vitesse tangentielle synchrone moyenne générale v_{mT} des points de l'induit disque, décalée d'un angle moyen θ_m par rapport à v_1 :

$$(11) \quad v_{mT} = \frac{v_1}{\cos \theta_m},$$

où

$$(12) \quad \cos \theta_m = \frac{2LR_m}{L \sqrt{\frac{L^2}{4} + R_m^2} + R_m^2 \ln \frac{L/2 + \sqrt{(L/2)^2 + R_m^2}}{-(L/2) + \sqrt{(L/2)^2 + R_m^2}}}$$

est le facteur moyen de décalage de vitesses. Dans cette relation R_m est le rayon correspondant au centre de symétrie des inducteurs et L —la longueur des inducteurs.

De la même manière on peut définir les composants moyens des forces déployées par le MLI : F_1 et respectivement F_{tr} , conformément aux composants des vitesses moyennes v_{mt} , respectivement v_{mtl} et v_{mtr} .

La force déployée par le MLI dans la direction tangentielle qui détermine un couple utile à l'axe de l'induit est donnée par la relation :

$$(13) \quad F_m = F_1 \cos \theta_m - F_{tr} \sin \theta_m.$$

En tenant compte de l'interdépendance de composants F_1 et F_{tr} de la force qui interviennent dans [7], on obtient :

$$(14) \quad F_1 = F_m K_f,$$

où :

$$(14) \quad K_f = \frac{1}{\cos \theta_m \left[1 - \frac{\tau v_{mt} \sin^2 \theta_m}{4l_1(v_1 - v_{mt} \cos \theta_m) \cos \theta_m} \right]}$$

est un coefficient de correction qui intervient lors de la détermination de la composante longitudinale F_1 de la force. À l'aide de cette relation on peut tracer la caractéristique mécanique $F_1 = f_1(v)$ du MLI, à l'induit en disque.

Dans les relations ci-dessus F_m a la forme indiquée par (9), à savoir $F_m = F_r$.

Nous avons déterminé les coefficients de correction selon la méthode de [7] basée sur certaines hypothèses simplificatrices, mais on entrevoit aussi une autre méthode d'estimation qui fera l'objet d'un travail subséquent.

5. Tests expérimentaux

On a testé divers moteurs linéaires sur le banc réalisé, en essayant de déterminer expérimentalement leurs caractéristiques mécaniques et leur comportement dans des divers régimes de fonctionnement (moteur, freinage par couplage inverse, freinage dynamique, etc.). Pour un MLI ayant les paramètres ci-dessous : $R_1=0,4773 \Omega$, $R_{20}=0,6028 \Omega$, $L_1=0,00803 \text{ H}$, $L_{m0}=0,0025 \text{ H}$, $2p=8$, $\tau_p=0,072 \text{ m}$, $f=50\text{Hz}$, $\Delta=0,003 \text{ m}$, $\delta=0,013 \text{ m}$, on présente, dans la fig. 3, la caractéristique mécanique calculée. De même, dans la fig. 4 on représente la variation du courant de phase au démarrage et, respectivement, la variation de la force de traction longitudinale en fonction du temps (les valeurs calculées sont représentées par une ligne continue, les valeurs mesurées expérimentalement par des points).

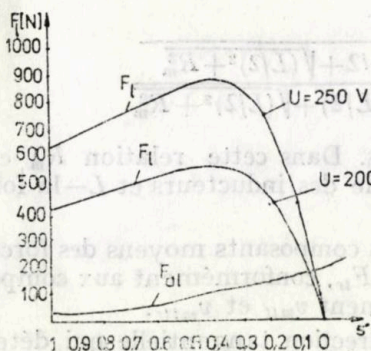


Fig. 3

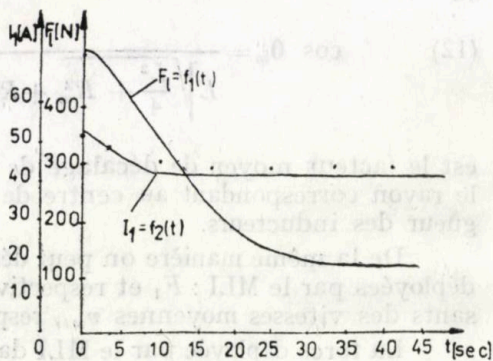


Fig. 4

6. Conclusions

On présente nombre de détails concernant un banc pour tester les moteurs linéaires d'induction.

On établit des relations de calcul qui permettent de déterminer la force longitudinale de traction déployée par le MLI. On introduit des coefficients de correction pour la vitesse synchrone et, respectivement, pour la force de traction qui tient compte du fait que les points de l'induit-disque en mouvement ont des vitesses synchrones différentes.

Il y a une bonne concordance entre les résultats du calcul et ceux obtenus par voie expérimentale.

Requie de 15 septembre 1983

Institut Polytechnique, Jassy,
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POSIBILITĂȚI DE ÎNCERCARE A MOTOARELOR LINIARE DE PUTERI MEDII

(Rezumat)

Se prezintă un stand original pentru încercarea motoarelor liniare de inducție de puteri medii, stabilindu-se relații de calcul pentru determinarea caracteristicilor mecanice și energetice.

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5495-5504, 5505-5514, 5515-5524, 5525-5534, 5535-5544, 5545-5554, 5555-5564, 5565-5574, 5575-5584, 5585-5594, 5595-5604, 5605-5614, 5615-5624, 5625-5634, 5635-5644, 5645-5654, 5655-5664, 5665-5674, 5675-5684, 5685-5694, 5695-5704, 5705-5714, 5715-5724, 5725-5734, 5735-5744, 5745-5754, 5755-5764, 5765-5774, 5775-5784, 5785-5794, 5795-5804, 5805-5814, 5815-5824, 5825-5834, 5835-5844, 5845-5854, 5855-5864, 5865-5874, 5875-5884, 5885-5894, 5895-5904, 5905-5914, 5915-5924, 5925-5934, 5935-5944, 5945-5954, 5955-5964, 5965-5974, 5975-5984, 5985-5994, 5995-6004, 6005-6014, 6015-6024, 6025-6034, 6035-6044, 6045-6054, 6055-6064, 6065-6074, 6075-6084, 6085-6094, 6095-6104, 6105-6114, 6115-6124, 6125-6134, 6135-6144, 6145-6154, 6155-6164, 6165-6174, 6175-6184, 6185-6194, 6195-6204, 6205-6214, 6215-6224, 6225-6234, 6235-6244, 6245-6254, 6255-6264, 6265-6274, 6275-6284, 6285-6294, 6295-6304, 6305-6314, 6315-6324, 6325-6334, 6335-6344, 6345-6354, 6355-6364, 6365-6374, 6375-6384, 6385-6394, 6395-6404, 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DYNAMISCHE PROGRAMMIERUNGSMODELLE FÜR DIE BESTIMMUNG DER OPTIMALZUVERLÄSSIGKEITSGRADEN FÜR DIE GLIEDER DER ELEKTROENERGIEVERSORGUNGSSYSTEM

VON

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1. Einleitung

Der von allen Fachmännern angenommene Sichtpunkt, daß für Optimierungsmodelle des Elektroenergieversorgungssystems (EEVS) auch der Zuverlässigkeitsgrad betrachtet werden muß, erscheint nicht mehr in der verwendeten Ausdrucksform für diese Betrachtungen. Es ist doch schon sehr klar, daß das von dem Typ von Abnehmer abhängt. Es gibt von diesem Gesichtspunkt aus folgende Abnehmerklasse:

a) Bei denen man Elektroenergieversorgungsunterbrechungsschäden (EEVUS) direkt in Geld ausgedrückt betrachten kann.

b) Bei denen diese Schäden durch die Gefahr für das menschliche Leben oder durch die Störungen des sozial-ökonomischen Lebens ausgedrückt werden können.

Im erstem Fall kann man die EEVUS ähnlich wie andere Aufwände (Investitionen, Betriebsaufwände) in ein ökonomisches Modell einführen. Dieses Modell kann die Zielfunktion für eine Optimierungsroutine sein. In dem zweitem Fall können die Zuverlässigkeitswerte des ganzen EEVS als Beschränkungswert in das Optimierungsmodell eingeführt werden. Sie müssen einem angenommenen Risikograd entsprechen. Das heißt, daß das Optimierungsmodell von der Abnehmerklasse abhängt.

Wie in [2] unterstrichen wird, sind die Optimierungsmodelle — in dem Falle der ersten Abnehmerklasse — noch von der Form der Schaden—Unterbrechungzeit-Kurve abhängig. Der Zuverlässigkeitswert für das ganze System hängt von den Zuverlässigkeitswerten der Gliedern und von der Systemstruktur (Schaltung) ab. Prinzipiell kann man das EEVS als eine Verkettung von mehreren Gliedern (Schaltanlagen, Leitungen, Transformatoren u.s.w.) betrachten. Diese Glieder können einzeln oder auch aus mehreren parallel geschalteten „Untergliedern“ gebaut sein. Abb. 1 versucht diese Untergliederung von EEVS für einen Abnehmer (C_1) darzustellen. Gleiche Zuverlässigkeitsversorgungsgrade für das ganze EEVS können mit verschiedenen und verschiedengeordneten Gliedertypen und Gliederzahlen erhalten werden. Davon kommen zwei verschiedene Fragen für die EEVS-optimisierung heraus. In dem Fall (a) ist zu finden sowohl der optimale Wert des Zuverlässigkeitsgrades für das ganze System als auch die Systemstruktur, welche diesen Wert des Zuverlässigkeitsgrades mit kleinstem Aufwand ausführen kann. In dem Fall (b) ist die Systemstruktur zu finden, die mit kleinstem Aufwand den Wert des Zuverlässigkeitsgrades versichert,

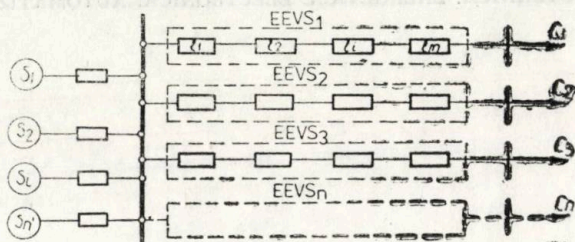


Abb. 1

der sich dem angenommenen Risiko anpaßt. Der zweite Fall ist leichter zu lösen und es gibt schon dafür wertvolle vorgeschlagene Lösungen [2]. Schwer zu lösen ist der erste Fall, besonders wenn man die komplexen Zuverlässigkeitswerte benutzt, wie Heufigkeits — Dauer-Unterbrechungsspektren und die Kurve Unterbrechungsdauer — EEVSUS. Solche Modelle sind fast immer unlösbar und in der ersten EEVS-Entwicklungsphase nicht nötig. So braucht man die leichtlösenden Modelle zu entwickeln, die auf einfachausdrückbare Zuverlässigkeitswerte wie Unterbrechungswahrscheinlichkeit (q) gegründet sind. Diese Modelle sollen nur die optimalen Strukturklassen für EEVS zeigen, um den Projektanten dahin zu führen. Weiter werden zwei solche Modelle entwickelt die von Verfasser schon vorgeschlagen waren [3], [4] und auf besonders behandelte [1] dynamische Programmierung nach Bellman gegründet sind. Um eine zweckmäßige Zielfunktion zu erzeugen, braucht man die Abhängigkeitsausdrücke zwischen Zuverlässigkeitsgradwert und verschiedene Kosten und Aufwände für alle Gliedertypen von EEVS. Diese Ausdrücke sind selbstverständlich verschieden für verschiedene Glieder, aber sie haben gemeinschaftlich, daß jede Steigung des Zuverlässigkeitsgrades desto mehr kostet, je größer der Zuverlässigkeitsgradwert wird. Die Form für diesen Ausdruck hängt von dem für die Zuverlässigkeitsgradsteigung angewandten Weg ab. Für die weiterentwickelten Modelle können diese Ausdrücke jedwelche Form annehmen (analytisch oder tabelarisch).

2. Entwicklung der Aufwände — Unterbrechungswahrscheinlichkeit- Abhängigkeitsausdrücke

Die Form für diese Abhangausdrücke hängt von dem verwendeten Weg für das Wachstum von Zuverlässigkeitswerten ab.

2.1. Der Fall wenn das Wachsen der Zuverlässigkeitswerte durch Reservierung (parallel geschaltete zusätzliche ähnliche Elemente) durchgeführt wird

Das ist der Fall von Leitungen, Transformatoren und anderen ähnlichen. Der Ausdruck des gesamten vergegenwärtigten Aufwandes kann die folgende Form annehmen [6]:

$$(1) \quad Z = K_i I + K_p p + K_d D,$$

wobei: K_i , K_p und K_d — Koeffizienten; I — Investitionskosten; p — Leistungsverluste; D — die Schäden, die an Versorgungsunterbrechungen mit Elektroenergie erscheinen.

Für eine Zahl n von parallel geschalteten Elemente geht für die gesuchte Funktion die folgende Form heraus:

$$(2.1) \quad I = n I_0, \quad (2.2) \quad q = q_0^n,$$

wobei: I_0 , I — Investitionskosten für ein Element bzw. für n Elemente; q_0 , q — Unterbrechungswahrscheinlichkeit für ein Element bzw. für n parallelgeschaltete Elemente.

Von der Ausdruck (2.1) geht durch Logarithmierung hervor:

$$(3) \quad \ln q = n \ln q_0$$

und dann:

$$(4) \quad n = \frac{\ln q}{\ln q_0}$$

und mit (2.1) zusammen:

$$I = \frac{I_0}{\ln q_0} \ln q.$$

Die Abhängigkeitsfunktion zwischen Leistungsverluste p und q geht von der Hauptfunktion für Leistungsverluste für n parallelgeschaltete Elemente heraus:

$$(6) \quad p = n p_0 + \frac{p_k}{n},$$

wobei: p_0 — Leerlaufverluste; p_k — Kurzschlußverluste, die mit der Funktion (4) zusammen gibt:

$$(7) \quad p = \frac{p_0}{\ln q_0} \ln q + \frac{p_k \ln q_0}{\ln q}.$$

Die Verbindung zwischen D und q ist schon bekannt:

$$(8) \quad D = q T d,$$

wobei: T — Rechnungszeit; d — Mittelwert der spezifischen Schäden für die Zeiteinheit.

2.2. Der Fall wenn das Wachsen der Zuverlässigkeitswerten durch Spaltungen der Netze durchgeführt wird

Das ist der Fall von Sammelschienen und doppelinspeisenden Netze. Die Form einer solchen Abhängigkeit ist im Abb. 2 dargestellt [5]. Hier, auf der Abszissenachse ist der Ausdruck $1 - q$ repräsentiert und auf der Ordinatenachse der spezifische Aufwand, auf eine Kilowattstunde gelieferte Energie bezogen. Diese Abhängigkeit kann man im EDVA tabella-

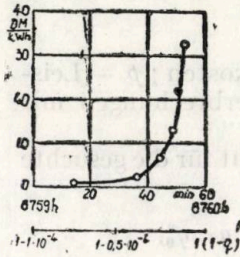


Abb. 2

angegeben sind. Zum Beispiel [7, S. 218] sind dieser Werte für die Zellen der verschiedenen möglichen Schaltungen für Schaltanlagen gegeben, die unmittelbar in den Programmen, die von den weiteren Modellen erscheinen, deklariert werden können. Ähnliche Werte kann man auch von [9] erzeugt wird.

risch oder analytisch deklarieren. Zum Beispiel ist der Aufwand für das Wachsen von q für die Schaltanlage durch Sammelschienenspaltung mit den Kosten der Kupplungszelle gleich.

2.3. Der Fall wenn die Aufwände und die Zuverlässigkeitswerte für verschiedene mögliche Lösungen bekannt sind

Das kann der Fall von Schaltanlagen sein, wo eine Formel wie die vorher gegebene schwer zu erzeugen ist. Es gibt aber mehrere gefertigte Entwürfe, wo alle Kosten (Aufwände) und Zuverlässigkeitswerte

3 Optimisierungsmodelle

Weiter werden die auf die vorher gefertigten Ausdrücke gegründeten zwei Optimisierungsmodelle gegeben.

3.1. Der gesamte (globale) Zuverlässigkeitsgradwert als Festwert (Beschränkung) auferlegt ist

Es sei ein EEVS, von m Serienelemente gebaut, wie im Abb. 3 angegeben wird. Dann wird der gesamte vergegenwärtigte Aufwand für das ganze System folgende Form annehmen :

$$(9) \quad Z = \sum_{j=1}^m (K_t I_j + K_p p_j).$$

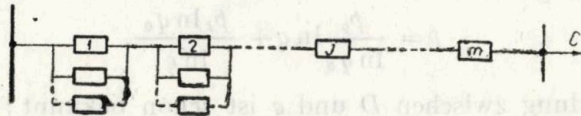


Abb. 3

Man muß unterstreichen, daß für einige Elemente, wie zum Beispiel Schaltanlagen, das zweite Glied keinen Sinn haben kann. Weiter wird trotzdem der globale Fall ausgearbeitet, weil, wenn ein Glied fällt, das Modell sich dann vereinfacht. Von den Ausdrücken (9), (5) und (7) geht heraus :

$$(10) \quad Z = \sum_{j=1}^m \left(K_t I_{0j} \frac{\ln q_j}{q_{0j}} + K_p \frac{p_{0j}}{\ln q_{0j}} \ln q_j + K_p p_k \frac{\ln q_{0j}}{\ln q_j} \right) = \sum_{j=1}^m \left(K_{1j} \ln q_j + \frac{K_{2j}}{\ln q_j} \right),$$

wo sich die Ausdrücke für K_{1j} und K_{2j} von selbst verstehen. Der Wert von q für das gesamte System, der in die Rechnung als Beschränkung eingeführt ist und mit einer Annäherung erster Ordnung (mit einem in diesem Fall annehmbaren Fehler von q^2 Ordnung) berechnet werden kann, wird folgenden Ausdruck haben :

$$(11) \quad q = \sum_{j=1}^m q_j.$$

Wenn man mit q_1 den Maximalen Wert für q kennzeichnet, dann nimmt das Optimierungsproblem folgende Form an :

$$(12) \quad \min \sum_{j=1}^m \left(K_{1j} \ln q_j + \frac{K_{2j}}{\ln q_j} \right), \quad \sum_{j=1}^m q_j < q_1, \quad 0 < q_j < 1.$$

Dieses Problem kann man leicht mit dem klassischen (Bellman) Algorithmus für dynamische Programmierung lösen.

3.2. Der gesamte Zuverlässigkeitsgradwert wie ein zusätzlicher Aufwand eingeführt ist

In diesem Fall haben die gesamten vergegenwärtigten Aufwände folgende Form :

$$(13) \quad Z = \sum_{j=1}^m \left(K_{1j} \ln q_j + \frac{K_{2j}}{\ln q_j} + K_{ad} q_j \right),$$

die minimisiert sein soll. Wenn das System schon von den anderen Bedingungen eine minimale unreservierte bedingte Struktur hat (die hier als natürliche Struktur bezeichnet wird), kann man sagen daß, das System einen natürlichen Wert (q_n) für q hat. Dieser Wert ist am größten während der ganzen Optimierungsrechnung ($q \leq q_n$). Das Optimierungsproblem wird folgende Form annehmen :

$$(14) \quad Z = \sum_{j=1}^m \left(K_{1j} \ln q_j + \frac{K_{2j}}{\ln q_j} + K_{ad} q_j \right) = \min, \quad \sum_{j=1}^m q_j \leq q_n, \quad 0 < q_j < 1.$$

Der Wert für q_n ist doch kein fester Wert, er ist nur der am größten mögliche Wert. Es kommt davon heraus, daß man das Minimum des Wertes für die Zielfunktion berechnen kann, wenn man schrittweise den dynamischen Programmierungsalgorithmus für immer kleinere Werte für $q < q_n$ anwendet (das heißt, daß die Rechnung gestoppt wird, wenn die Zielfunktion wieder zu steigen beginnen wird). Die ganze Rechnung wird alle optimale Werte für q_j geben. Aus diesen Werten kann man den optimalen Reservierungsgrad (die optimale Zahl für die ähnlichen parallel geschalteten Elemente) berechnen. Das kann man sowohl mit der Gleichung :

$$(15) \quad n_{\text{opt.}} = \frac{\ln q_{\text{opt.}}}{\ln q_b}$$

als auch durch unmittelbare Lesung von der Tabelle (wenn die vorher behandelte Funktion tabellarisch deklariert ist) berechnen.

4. Schlussfolgerungen

Die Methoden für die EEVS-optimisierung müssen immer der Annehmerklasse angepaßt werden. Das heißt, daß man die Rechnungsmodelle verarbeiten muß, die sowohl immer mehr richtiger als auch lösbar sein sollen. In dieser Arbeit sind zwei solcher Optimierungsmodelle vorgeschlagen die : a) den dynamischen Programmierungsalgorithmus benutzen ; b) alle Aufwändeglieder eingeführen ; c) erlauben daß die Abhängigkeitsausdrücke zwischen den Zuverlässigkeitswerten und den verschiedenen Kosten sowohl analytisch als auch tabellarisch deklariert werden können.

Es sind auch verschiedene Wege gegeben, um diese Abhängigkeitsausdrücke zu erzeugen.

Eingegangen am 24. März 1983

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DETERMINAREA PRIN PROGRAMARE DINAMICĂ A NIVELELOR OPTIME DE FIABILITATE A COMPONENTELOR SISTEMELOR DE ALIMENTARE CU ENERGIE ELECTRICĂ

(Rezumat)

Modelele propuse sînt generalizări ale unor lucrări anterioare ale autorului și au la bază dependența dintre costul instalațiilor și indicatorilor lor de fiabilitate.

Se prezintă căile de obținere a acestor dependențe, modul de introducere a lor în modele precum și căile de rezolvare a modelelor rezultate

PROGRAMME POUR LE CALCUL DES PERTES D'ÉNERGIE DANS LES RÉSEAUX ÉLECTRIQUES

PAR

GH. CÎRȚINĂ, GH. HAZI et EUGEN TOMOZEI

1. Considérations générales

Dans un travail précédent [1] on a montré que par l'utilisation des caractéristiques statistiques (la moyenne, la variance et la covariance) des tensions $V(t)$ et des puissances aux noeuds $P(t)$, $Q(t)$, par le développement en série Taylor des pertes d'énergie, on arrive à des modèles mathématiques qui évitent le calcul successif des régimes et qui s'avèrent efficaces surtout dans les problèmes d'optimisation.

Dans ce qui suit on présente un programme général de calcul, ses sous-routines et la manière d'utilisation.

Dans un réseau maillé avec $n+1=N$ noeuds (N - le noeud d'équilibre) les pertes de puissance ont l'expression [2]:

$$\Delta P = \sum_{i=1}^n \sum_{j=1}^n \frac{1}{V_i V_j} [(P_i P_j + Q_i Q_j) R_{ij} \cos \delta_{ij} + (Q_i P_j - P_i Q_j) R_{ij} \sin \delta_{ij}] + \\ + V_e \sum_{i=1}^n \frac{C_{ei}}{V_i} (P_i \sin \delta_i - Q_i \cos \delta_i) + G_e V_e^2. \quad (1)$$

où :

$$\underline{V}_i = V_i \underline{\delta}_i, \quad (i = \overline{1, n}); \quad (2) \quad \underline{V}_e = V_e \underline{0}; \quad (3)$$

$$R_{ij} = \mathcal{R}_e(\underline{Z}_{ij}) - \text{la partie réelle de l'élément } \underline{Z}_{ij} \text{ de la matrice } [\underline{Z}_{nn}]; \quad (4)$$

$$[\underline{Z}_{nn}] = [\underline{Y}_{nn}]^{-1} \quad (5)$$

$$[\underline{Y}_{NN}] = \begin{bmatrix} [\underline{Y}_{nn}] & -[\underline{Y}_{ne}] \\ -[\underline{Y}_{en}] & [\underline{Y}_{ee}] \end{bmatrix} - \text{la matrice des admittances des noeuds}; \quad (6)$$

$$G_e = \mathcal{R}_e[\underline{Y}_e] = \mathcal{R}_e\{[\underline{Y}_{ee}] - [\underline{Y}_{en}] [\underline{Z}_{nn}] [\underline{Y}_{ne}]\}; \quad (7)$$

$$C_{ei} = \mathcal{I}m \left\{ \sum_{j=1}^n (\underline{Y}_{ej} \underline{Z}_{ji} - \underline{Y}_{ej}^* \underline{Z}_{ji}) \right\}. \quad (8)$$

Par développement en série Taylor de la fonction $\Delta P(P, Q, V_e)$ aux environs des valeurs moyennes $\bar{P}_i, \bar{Q}_i, (i = \overline{1, n})$, et en calculant la

moyenne en temps on obtient [1] :

$$(9) \quad \overline{\Delta P}_{PQ} = \Delta P(\bar{P}, \bar{Q}, \bar{V}_e) + C\emptyset RPQ.$$

Alors, les pertes d'énergie dans la période T auront pour expression

$$(10) \quad \Delta W = \overline{\Delta P}_{PQ} T.$$

2. Le programme principal de calcul

Le schéma de calcul est donné, en principe, dans la fig. 1. Dans la fig. 2 est présenté le schéma du programme segmenté.

Le programme principal contient les instructions :

```

1  REAL *8 TIT(6)
2  1  READ(105, 10, END= 999) N, M, NATR, NIT, EPS, (TIT(I), I= 1, 6), IS, IP
3  10  FORMAT(4I5, F10.5, 6A8, 2I1)
4  WRITE(108, 11) (TIT(I), I= 1, 6)
5  11  FORMAT(126('')/2H'', 39X, 6A8, 36X, 1H*/126(''))
6  NN= N- 1
7  CALL STAT(NN, M, IS)
8  CALL FØRMY(NN, NATR)
9  CALL INVMAT(NN)
10 CALL REGPER(NN, NIT, EPS, IP)
11 CALL C ØR(NN)
12   G Ø T Ø 1
13 999 ST ØP
14   END

```

où : N — le nombre des noeuds du réseau ; M — le nombre des paliers des courbes de charge ; $NATR$ — le nombre des types des transformateurs ; NIT — le nombre d'itérations admis dans le calcul du régime stationnaire ; EPS — l'erreur (en kV) admise dans le calcul du régime stationnaire ; IS — l'indicateur d'impression des caractéristiques statistiques ; IP — l'indicateur d'impression des caractéristiques du régime stationnaire.

3. Les sousprogrammes de calcul

Le programme général contient 5 sousroutines : $STAT$, $FØRMY$, $INVMAT$, $REGPER$ et $CØR$.

3.1. Subroutine STAS (NN, M, IS)

Cette subroutine calcule les caractéristiques statistiques pour les tensions $V(t)$ et les puissances aux noeuds $P(t)$, $Q(t)$ en utilisant les relations :

a) pour la moyenne

$$(11) \quad \bar{X} = XM = \frac{1}{M} \sum_{i=1}^M X_i;$$

b) pour la variance

$$(12) \quad \sigma_x^2 = \text{SIGMAX} = \frac{1}{M} \sum_{i=1}^M (X_i - XM)^2;$$

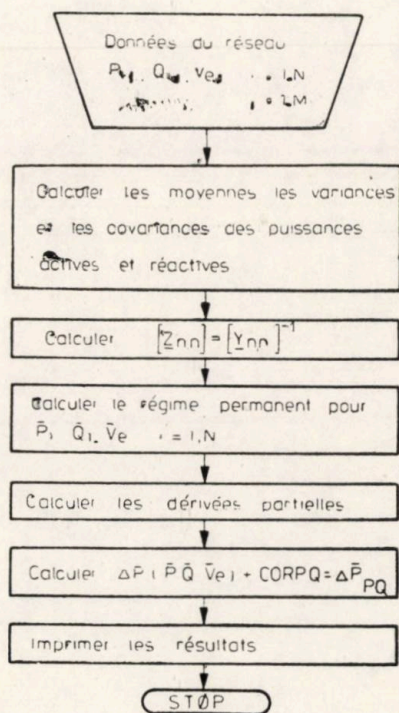


Fig. 1

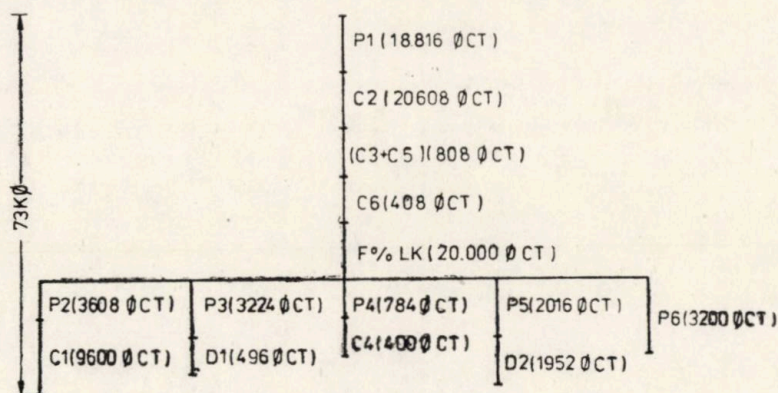


Fig. 2

c) pour la covariance

$$(13) \quad \text{Cov}(X, Y) = \frac{1}{M} \sum_{i=1}^M (X_i - XM)(Y_i - YM).$$

3.2. Subroutine FORMY (NN, NATR)

Dans cette subroutine on lit les données des transformateurs et des lignes, on calcule les paramètres correspondants et on forme la matrice $[Y_{nn}]$, dont les éléments sont :

$$(14) \quad -Y_{ik} = \sum_{j=1}^l a_{ij} Y_j a_{kj}^*, \quad (i \neq k); \quad (15) \quad Y_{ii} = \sum_{j=1}^l a_{ij}^2 Y_j,$$

avec l — le nombre des cordes du graphe du réseau, a_{ij} — l'élément (i, j) de la matrice d'incidence cordes — noeuds.

Ces relations permettent à écrire directement la matrice $[Y_{nn}]$. Les admittances Y_{ii} , $(i = \overline{1, n})$, sont calculées comme la somme des admittances incidentes au noeud i . Si la corde contient un transformateur, l'admittance de la corde respective doit être multipliée par le carré du rapport de transformation.

3.3. Subroutine INVMAT(NN)

Cette subroutine effectue l'inversion de la matrice des admittances aux noeuds. Ses instructions sont :

1	SUBROUTINE INVMAT(NN)	9	DØ 420 J=1,NN
2	CØMPLEX Y(50,50),PIV(50)	10	Y(I,J)=Y(I,J)/PIV(I)
3	CØMM ON Y/C4/PIV	11	DØ 420 L=1,NN
4	DØ 420 I=1,NN	12	IF(L.E.Q.I) GØ TØ 420
5	DØ 410 J=1,NN	13	Y(L,J)=Y(L,J)-Y(I,J)*PIV(L)
6	PIV(J)=Y(J,I)	14	420 CØNTINUE
7	410 Y(J,I)=(.0,.0)	15	RETURN
8	Y(I,I)=(1...0)	16	END

L'inversion de la matrice $[Y_{nn}]$ est effectuée par la méthode du pivot et elle est basée sur le fait que cette matrice est nonsingulière, les éléments de la diagonale sont différents de zéro et, généralement, plus grands que les éléments nondiagonales.

3.4. Subroutine REGPER (NN, NIT, EPS, IP)

Par l'appellation de cette subroutine on calcule le régime stationnaire pour les puissances moyennes (aux noeuds du réseau), à l'aide de la méthode des potentiels aux noeuds avec les relations :

$$(16) \quad \underline{V}_i = \underline{V}_{ni}, \quad (i = \overline{1, n}),$$

$$(17) \quad \underline{J}_i = \frac{S_i^*}{\underline{V}_i^*}, \quad (i = \overline{1, n}),$$

$$(18) \quad [\underline{V}_n] = [\underline{Z}_{nn}] [\underline{J}_n] + [\underline{Z}_{nn}] [\underline{Y}_{ne}] \underline{V}_e.$$

Le nombre des itérations est généralement 3...4, pour $EPS=0,02$ kV et 6...9 pour $EPS=0,01$ kV.

3.5. Subroutine CØR (NN)

Ce programme permet à calculer les pertes de puissance $\overline{\Delta P}_{PQ}$, $\Delta P(\overline{P}, \overline{Q}, V_e)$ et la correction en utilisant les relations [1] :

$$(19) \quad \Delta P(\overline{P}, \overline{Q}, \overline{V}_e) = V_e^2 G_{ee} - V_e \sum_{i=1}^n V_i (G_{ie} \cos \delta_i + B_{ie} \sin \delta_i) + \sum_{i=1}^n P_i,$$

$$(20) \quad \begin{aligned} \text{CORPQ} = & 0,5 \sum_{i=1}^n \sum_{j=1}^n \left[\left(\frac{\partial^2 \Delta P}{\partial P_i \partial P_j} \right)_m \text{Cov}(P_i, P_j) + \right. \\ & + \left(\frac{\partial^2 \Delta P}{\partial Q_i \partial Q_j} \right)_m \text{Cov}(Q_i, Q_j) + 2 \left(\frac{\partial^2 \Delta P}{\partial P_i \partial Q_j} \right)_m \text{Cov}(P_i, Q_j) \Big] + \\ & + \sum_{i=1}^n \left[\left(\frac{\partial^2 \Delta P}{\partial P_i \partial V_e} \right)_m \text{Cov}(P_i, V_e) + \left(\frac{\partial^2 \Delta P}{\partial Q_i \partial V_e} \right)_m \text{Cov}(Q_i, V_e) \right] + \\ & + 0,5 \left(\frac{\partial^2 \Delta P}{\partial V_e^2} \right)_m \text{Cov}(V_e, V_e). \end{aligned}$$

4. Conclusions

Le programme a été testé pour différents réseaux, avec un ou plusieurs plans de tensions, avec un nombre de noeuds allant de 5 à 50, un nombre de paliers de la courbe de charge de 4 à 48 et des variations plus ou moins conformes des charges.

Le programme permet une réduction substantielle du temps de calcul (8...10 fois).

Le modèle et le programme peuvent être utilisés pour la minimisation de l'énergie perdue.

Reçu le 1er juin 1983

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PROGRAM PENTRU CALCULUL PIERDERILOR DE ENERGIE ÎN REȚELELE ELECTRICE

(Rezumat)

Se prezintă un program general și principalele subrutine pentru calculul pierderilor de energie utilizând caracteristicile statistice ale sarcinilor. Programul permite o reducere substanțială a timpului de calcul.

EINIGE BEMERKUNGEN ÜBER DEN MAGNETISCHEN FREQUENZVERDOPPLER IN DER AUSSICHT DER VERWENDUNG ALS SPEISEQUELLE FÜR PRÜFUNG MIT INDUZIERTER WECHSELSPANNUNG DER ELEKTRISCHEN ISOLATION VON SPANNUNGSWANDLERN

VON

RODICA OLĂH und N. GAVRILĂȘ

1. Vorstellung des Problems

In der Nomenklatur der Prüfung der elektrischen Isolation nimmt die Prüfung mit vergrößerter induzierter Wechselspannung der Isolation von Leistungstransformatoren und Spannungswandlern einen wichtigen Platz ein. Im Rahmen dieser Prüfungen ist eine Vergrößerung der Speisespannungsfrequenz notwendig, über den nominalen Wert von 50 Hz, für das Vermeiden der Magnetisierungsströme welche zu groß sind, verursacht von dem hohen Wert der Prüfspannung im Vergleich zur Nominalspannung [1], [2]. Die Frequenzdrehkonvertere, in diesem Zweck verwendbar, haben den Nachteil einer relativ komplizierten Ausführungstechnologie, welche sie unökonomisch macht und zwar: a) das Vorhandensein von Schwingungen und Lärm während der Funktion, welche die Feststellung der Defekte in der Isolation durch Hören erschwert; b) Abnützung und schwere Instandhaltung.

Im Sinne der Beseitigung dieser Nachteile und für die Erfüllung der praktischen Notwendigkeit daß man bei Reparaturarbeiten und/oder Modernisierung von Spannungswandlern (SW) über eine kräftige Prüfausrüstung verfügt und mit kleineren Investitionen zu realisieren¹⁾ haben die Verfasser verschi dene Funktionsansichten der magnetischen Frequenzverdoppler versucht in der Aussicht der Verwendung dieser als Speissspannungsquellen der Schaltungen der Prüftechnik mit induzierter Wechselspannung für die Isolation von SW.

2. Erforderliche technische Bedingungen

Den Speisequellen aus den Prüfausrüstungen der Isolation mit induzierter Wechselspannung werden folgende wichtige technische Bedingungen gestellt: a) sie sollen die induktive Last, die durch die spezifischen Prüfschaltungen (meistens einphasische) des geprobten Transformators dargestellt ist, bei der vergrößerten Spannungsfrequenz im Verhältnis zu 50 Hz einspeisen; b) stufenlose Veränderung der Spannung mit vergrößerter Frequenz, bis zu dem betreffenden Wert der Prüfspannung, die Spannungssprünge vermeidend welche instationäre Regime zur Folge hätten; c) der Kurzschlußstrom der Quelle soll genügend groß sein damit

¹⁾ Die erforderliche Arbeit wurde von Ministerium für Elektroenergie mittels URTAE Roman - Werk verlangt [3].

im inkompleten Entladungsregime nicht eine Begrenzung der Spannung erscheint; d) die Kurvenform der Spannung bei der vergrößerten Frequenz soll so ähnlich wie möglich mit der Sinusoide sein (charakterisiert durch eine Abweichung des Scheitelfaktors im Vergleich zu $\sqrt{2}$, von maximum $\pm 5\%$); e) es ist nötig daß sie den Magnetisierungsstrom vom erprobten Transformator, bei vorgeschriebener Prüfungsspannung, begrenzen; f) der Wert der Frequenz soll die Gültigkeit der Prüfergebnisse nicht beeinflussen und soll der Kostenpreis der Prüfeinrichtung nicht zunehmen; g) sie sollen ein gutes dynamisches Verhalten im Übergangsbereich zwischen zwei stationären Zuständen haben.

Die Gesamtheit der oben erwähnten Bedingungen führt zu dem einphasigen Schema des Frequenzverdopplers (DMF) das vom Drehstromnetz gespeist wird (Abb. 1). In Bezug auf dieses Schema werden mit Index 1

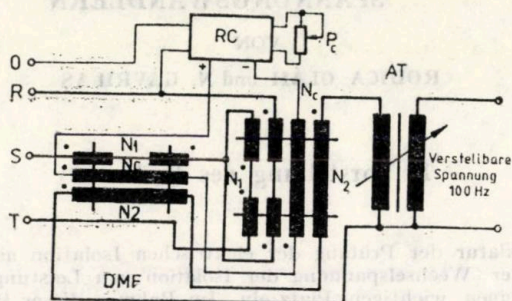


Abb. 1

die Größen auf den Primär, mit 2 jene die sich auf den Sekundär beziehen und mit c , die Gleichstromvormagnetisierungswicklung, welche vom gesteuerten Gleichrichter RC gespeist wird. In diesem Schema AT ist ein verstellbarer Transformator welcher die Änderung der Spannung mit vergrößerter Frequenz ermöglicht.

Weiterhin werden einige Funktionsansichten des magnetischen Frequenzverdopplers, in Verbindung mit den durch die angenommene Verwendung gelegte Bedingungen hervorgehoben.

3. Experimentaluntersuchung über das Verhalten des magnetischen Frequenzverdopplers als Speisequelle für die Prüfung mit induzierter Wechselspannung der Isolation von Spannungswandlern

Die Versuche wurden mit einer experimentellen Installation, die von den Verfassern, nach dem Prinzipschema aus Abb. 1, entworfen und erzeugt wurde, welche eine Nennleistung von 3,5 kVA hat und bei welcher das Transformatorenblech vom Typ MGX von 0,8 mm verwendet wurde, durchgeführt.

Es wurde das Auswirken des Gleichstrommagnetfeldes H_c und des Maximalwertes der Induktion B_{1m} der Grundwelle auf den Wert und auf die Kurvenform der Leerlaufspannung U_{20} , sowie auf den Wert des Sekundärkurzschlußstromes I_{2sc} , verfolgt. Zu diesem Zweck wurden die Abhängigkeiten $U_{20}=f(H_c)$ und $I_{2sc}=f(H_c)$, beide für $B_{1m}=\text{const.}$ (Abb. 2) wie, auch

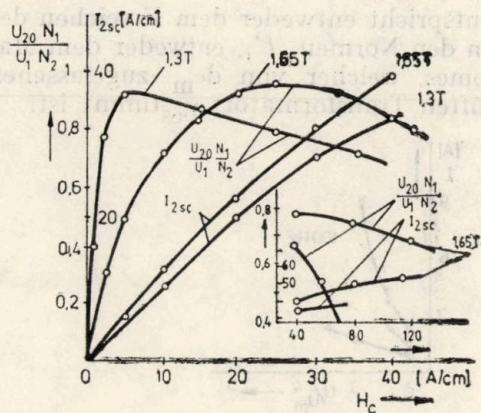


Abb. 2

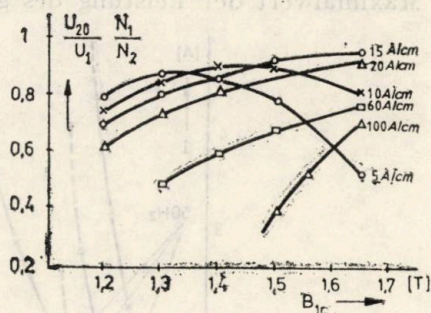


Abb. 3

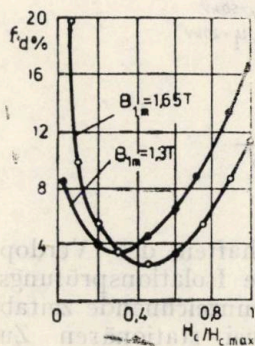


Abb. 4

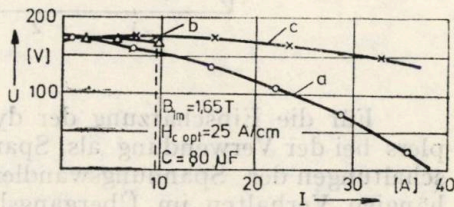


Abb. 5

$U_{20} = f(B_{1m})$ für $H_c = \text{const.}$ (Abb. 3) dargestellt²⁾. Die Art in welcher die Kurvenform der Spannung U_{20} von den Werten H_c und B_{1m} beeinflusst ist, ist durch die Abhängigkeit des Formfaktors wiedergegeben, ($f_a\% = f(H_c)$) für $B_{1m} = \text{const.}$, aus Abb. 4, erhalten durch Verarbeitung einer großen Menge von Oszillogrammen. Abb. 5 wiedergibt die statische Spannung—Stromkennlinie $U_2 = f(I_2)$ für $H_c = \text{const.}$, bei aktiver Belastung (Kurve a), bei induktiver Belastung, dargestellt von einem Spannungsmeßtransformator TEMU-110 (Kurve b) und die erstarten Kennlinien bei aktiver Belastung mit einer longitudinalen kapazitiven Kompensation (Kurve c) erhalten.

Der Effekt der Frequenzverdopplung auf die Prüfmöglichkeiten der elektrischen Isolation wird im Abb. 6 durch die Abhängigkeit des aufgenommenen Stromes von der Änderungsspeisespannung $I = f(U)$ wiedergegeben, bei 50 Hz und bzw. 100 Hz, in Falle mancher Spannungswandler. Der

²⁾ Für die Verallgemeinerung der Ergebnisse wurde die Sekundärspannung durch das Verhältnis $U_{20} = N_2 / (U_1 N_1)$ und der Sekundärkurzschlußstrom durch den Effektivwert des betreffenden Magnetfeldes ersetzt.

größte wert des Verhältnis³⁾ U/U_m entspricht entweder dem Erreichen der Nominalprüfspannung, gefordert von den Normen, U_t , entweder dem Maximalwert des aufgenommenen Stromes, welcher von dem zugelassenen Maximalwert der Leistung des geprüften Transformators bestimmt ist.

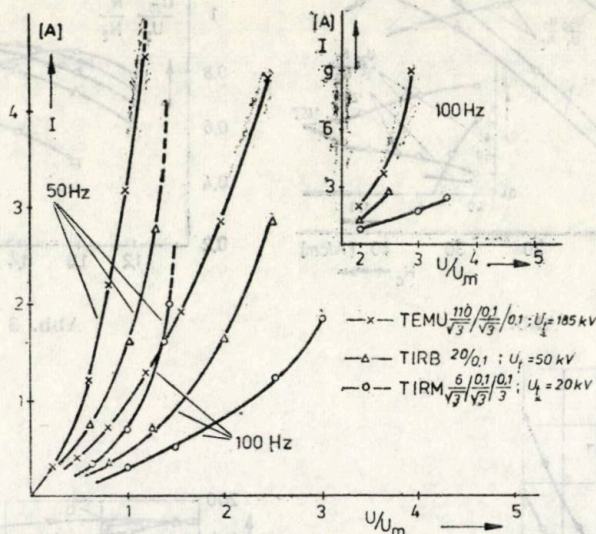


Abb. 6

Für die Einschätzung der dynamischen Eigenschaften des Verdopplers bei der Verwendung als Spannungsquelle für die Isolationsprüfungsschaltungen der Spannungswandler wurden einige kennzeichnende zeitabhängige Verhalten im Übergangsbereich zwischen zwei stationären Zuständen beobachtet. Abb. 7 a bezieht sich auf den Übergangsvorgänge zwischen Leerlauf und Kurzschluß und Abb. 7 b zwischen Kurzschluß und Leerlauf. Diese Oszillogramme sind für die Werte des Stationärbetriebs dargestellt worden ($U_{20} = 180$ V, $I_{200} = 40$ A).

Der zeitliche Verlauf der Spannung und des Stromes während eines Isolationsdurchschlagversuches für einen TEMU-110 KV ist im Abb. 7 c für den Fall eines schroffen und im Abb. 7 b für einem langsamen Durchlag widergegeben. Die Oszillogramme wurden an den Sekundärwicklungsklemmen des geprüften Spannungswandlers aufgezeichnet, bei einer Speisespannung von $U = 169$ V, entsprechend der Nominalprüfspannung $U_t = 185$ KV der Wicklung mit ungleichrepartierter Isolation von 110 KV.

4. Deuten der Ergebnisse und Schlußfolgerungen

Aus den durchgeführten Versuche ergeben sich folgende Bemerkungen:

1. Der Effektivwert der Leerlaufsekundärspannung des Frequenzverdopplers welcher das Prinzipialschema in Abb. 1 und eine schwach

³⁾ U_m ist die Maximalbetriebspannung der Wicklung.

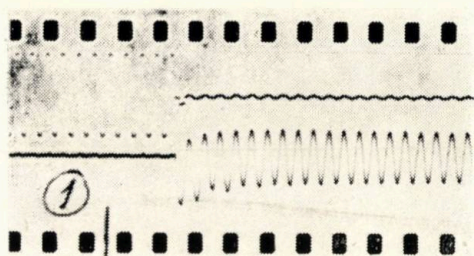


Abb. 7 a

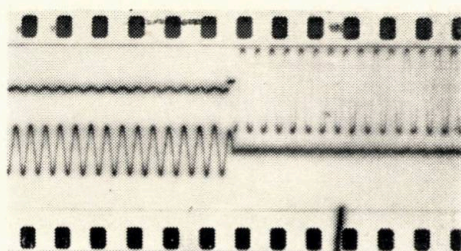


Abb. 7 b

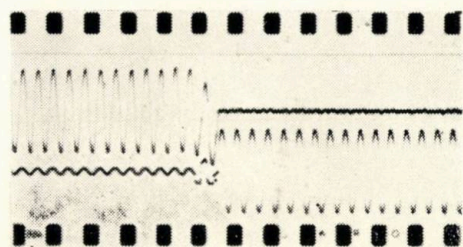


Abb. 7 c

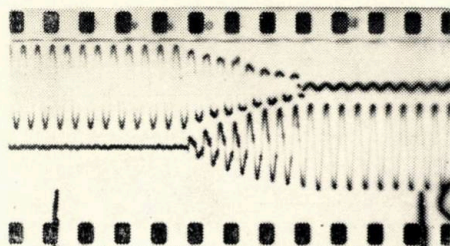
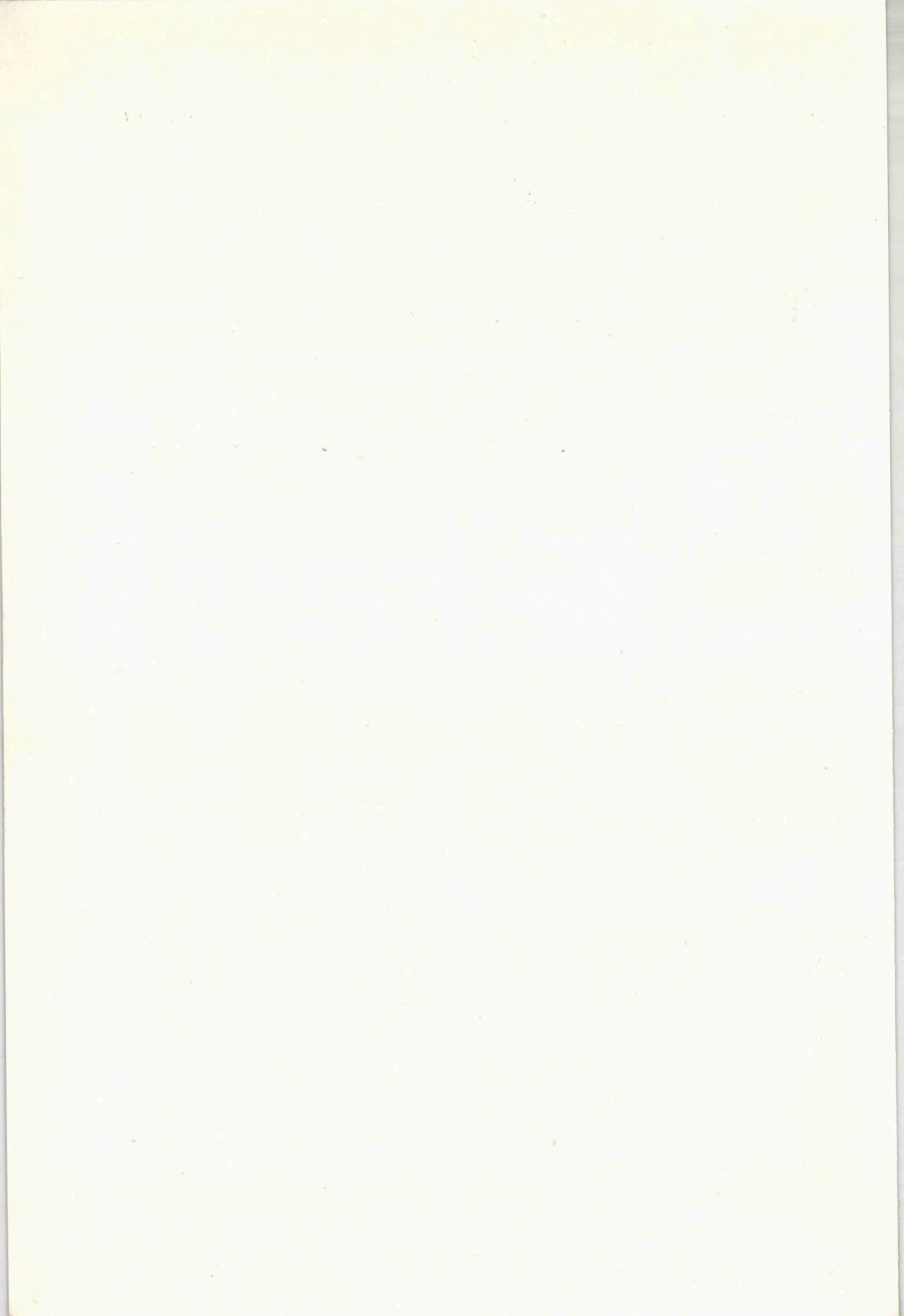


Abb. 7 d



unliniare Magnetisierkennlinie (das Blech M6X von 0,3 mm) hat, ist von dem Maximalwert der Grundwelleninduktion B_{1m} , bestimmt von der Primärspannung U_1 , wie auch von der Intensität des Gleichstrommagnetfeldes H_c , abhängig.

2. Der Maximalwert der Leerlaufsekundärspannung bei einem gegebenen Verhältnis N_2/N_1 , kann durch die entsprechende Zusammensetzung der Werte H_c und B_{1m} gesichert werden. Für $B_{1m}=1,3$ T hat sich der Optimalwert der Intensität des Gleichstrommagnetfeldes $H_{c, \text{opt}}=5$ A/cm ergeben und für $B_{1m}=1,65$ T ist $H_{c, \text{opt}}=25$ A/cm.

3. Eine Verkleinerung der Effektivleerlaufsekundärspannung pro Wicklung entspricht mit dem Wachsen des Oberschwingungengehaltes der Ausgangsspannungskurvenform des Frequenzverdopplers.

4. Die Optimalform der Ausgangsspannungskurve, nahe der Sinusoidenform, welche der Bedingung d) (§ 2) entspricht, gefordert von der Prüftechnik, kann man bei einem gegebenen Wert B_{1m} , für $H_{c, \text{opt}}$ erhalten oder höchstens in einem engen Wertbereich der Größe H_c in Nahe des Optimalwertes.

5. Für die gleichzeitige Erfüllung der Bedingungen b) und d) (s. § 2) muß der Frequenzverdoppler als unverstellbare Spannungsquelle arbeiten. Die Stufenlose Veränderung der Spannung mit der vergrößerten Frequenz, nötig für die Isolationsprüfung, muß im Sekundär des Frequenzverdopplers erfolgen, mittels eines einphasischen verstellbaren Transformators.

6. In der Aussicht der Erfüllung der Bedingung c), (§ 2) daß man eine Sekundärspannungsbegrenzung im Leistungslauf des Frequenzverdopplers vermeidet, ist eine Konstantspannungscharakteristik nötig. Im Rahmen dieser Verwendung kann man ein Verhältnis zwischen dem Sekundärkurzschluß- und Leistungsstrom des Frequenzverdopplers $I_{2sc}/I=5...6$ erhalten, weil der Leistungsstrom I durch den vom geprüften Spannungswandler aufgenommenen Magnetisierungsstrom dargestellt wird.

7. Wenn man größere Werte für H_c und B_{1m} (wie auch dem Sättigungswerte nahegelegene) annimmt so erhält man einen hinreichenden Wert für I_{2sc} . Für das Blech M6 von 0,3 mm ergibt sich die obere Grenze $H_{c, \text{max}}=120$ A/cm für die Sättigungswerte der Grundwelleninduktion $B_s=1,65$ T.

8. Wenn das Verhältnis I_{2sc}/I hinreichend groß ist, dann ist eine Erstarrung der statischen Kennlinien des Frequenzverdopplers mit einer longitudinalen kapazitiven Kompensation nicht nötig. Außerdem konnte das Vorhandensein dieser bei induktiver Leitung, dargestellt durch den geprüften SW, zu Unstabilitäterscheinungen führen.

9. Das dynamische Verhalten des Frequenzverdopplers bei plötzlichen Veränderungen der Leistung, welche bei der Isolationsprüfung auftreten können, wenn diese durchschlagen wird, ist günstig. Der Spitzenwertkurzschlußstrom klingt exponentiell zum Dauerkurzschlußstrom ab in einem Zeitintervall von höchstens 20...30 ms, das Verhältnis $i_{sc, \text{max}}/\sqrt{2} I_{sc}$ hat einen Wert von ungefähr 1,25 und auch keine Spannungsspitzenwerte treten auf.

10. Die durchgeführten Untersuchungen bei verschiedenen Spannungswandlern zeigt daß bei der Frequenz von 1.0 Hz, im Rahmen der Prüfung

mit induzierten Wechselspannung, der Prüfspannungswerte von bis zu $3,2 U_m$ erreichen kann, ohne daß der Wert des aufgenommenen Stromes den entsprechenden zugelassenen Maximalwert der Leistung des geprüften Transformators erreicht. Andererseits ist eine Prüfspannung von $3,2 U_m$, in Bezug auf die Beanspruchung der Isolation der Spannungswandler unter Betriebsbedingungen ausreichend [3]. In dieser Weise kann die Longitudinalisolation aller Spannungswandlern wie auch die Hauptisolation der meisten Typen mit nichtuniformer Isolation (bei welchen, daß Verhältnis U_i/U_m kleiner als 3,2 ist) bei einer von den Normen erforderten Prüfwechselspannung geprüft werden.

In Schlußfolgerung ist das Entwerfen und Verwirklichen einer Spannungsquelle mit Ausgangsfrequenz von 100 Hz möglich, welche auf dem Prinzip des magnetischen Frequenzmultiplikators beruht und welche der Isolationsprüfung der Spannungswandlern mit vergrößerter induzierter Wechselspannung, in erster Reihe im Rahmen der Prüfstände des elektroenergetischen Systems bestimmt ist.

Eingegangen am 23. Juni 1983

Technische Hochschule, Jassy,
Lehrstuhl für Elektroenergetik

LITERATURVERZEICHNIS

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2. Cristescu D., Oláh R., *Supratensiuni și izolația rețelelor electrice*. EDP., Buc., 1983.
3. Oláh R., Oláh I., *Instalație pentru încercarea izolației cu tensiune alternativă de frecvență mărită*. Forschungstrag, 1983.

OBSERVAȚII ASUPRA DUBLORULUI MAGNETIC DE FRECVENȚĂ ÎN VEDEREA UTILIZĂRII CA SURSĂ PENTRU ÎNCERCAREA CU TENSIUNE ALTERNATIVĂ INDUSĂ A IZOLAȚIEI TRANSFORMATOARELOR DE TENSIUNE

(Rezumat)

Se efectuează un studiu experimental al dublorului magnetic de frecvență în corelație cu cerințele tehnicii încercării izolației electrice, în scopul realizării unor instalații statice pentru încercarea cu tensiune alternativă indusă a izolației transformatoarelor de tensiune.

RECENZII

BOOK REVIEWS

РЕЦЕНЗИИ

MORGAN D. P., *Surface Accoustic Wave Passive Interdigital Devices*. Peter Peregrinus Ltd., 1976, 448 pp.

Surface acoustic waves (SAW-s) have received considerable attention recently, many SAW devices being now components of electronic systems. The use of the interdigital transducer for generation and reception of SAW-s determined in the last ten years an enormous interest about a phenomenon discovered in the 19-th century — the Rayleigh wave.

The book is a collection of selected reprinted articles concerning properties and applications of SAW-s in passive devices.

The first two papers give an introduction to interdigital devices and a review of their applications. Section 2 contains principles and properties of direct piezoelectric coupling with interdigital transducers. In Section 3 papers on materials and propagation effects concentrates on topics relevant to current devices.

In Section 4 on bandpass filters principles, capabilities and applications of SAW filters both with techniques for improving performances are discussed. Section 5 is concerned with two types of matched filter — dispersive filters for pulse compression and filters with discrete phase coding while Section 6 covers miscellaneous techniques mainly for improving device performances. At last, Section 7 contains an extensive thematic bibliography.

The book can be considered as a reference source of a real use not only for researcher working in the subject but also for readers new to the subject, with a more general interest.

Finally we conclude that the work of the editor, professor David P. Morgan, well-known researcher in the field, is an absolute success, the book being a remarkable contribution.

L. Goraş

BRODSKY M. H. (ed.), *Amorphous Semiconductors*. Serie *Topics in Applied Physics*, vol. 36, Springer Verlag, Berlin, 1979, 337 pp.

This book, edited by Marc H. Brodsky, who writes also the Preface and Introduction, is a collection of ten articles on the main topics in amorphous semiconductors.

As stated in the Preface, „volume addresses the physics of electronic phenomena in amorphous semiconductors, with emphasis principally dependent upon energy levels in and near the semiconductor band gap“.

The structure of the book follows the way from basics to applications.

In the Introduction, M. H. Brodsky defines the amorphous semiconductors, discusses the current and future trends and presents the organisation of the book.

The theory of electric states in amorphous semiconductors with reference to conductivity and critical behaviour is developed by B. Kramer and D. Weaire in the second chapter, while the states in the gap and defects, with discussions on the distinction between

of Ge - type and Si - type materials, and with reference to Ge, Si, chalcogenides and a-Arsenic are exposed by E.A. Davis in the third chapter.

Two outstanding chapters, dealing with the optical properties and with electronic transport in amorphous semiconductors are written by G.A.N. Connoll and respectively by P. Nagels.

Luminescence, spin effects and short-range order in amorphous semiconductors are treated by R. Fischer, I. Solomon, G. Lucovsky and T. M. Hayes in chapters 6, 7, and 8 respectively.

The last two chapters, *Doped Amorphous Semiconductors* and *Amorphous Silicon Solar Cells*, written by D. E. Carlson and C. R. Wronski, are applications oriented.

An annex with additional up-dated (1979) general references and a subject index ends the book.

It is obvious that the editor succeeded in giving coherence and continuity to the book, despite the large area of the topics developed.

H. N. Teodorescu

THIBERVILLE L., CHAPPERT R., CAMPA A., *L'automatique par les problèmes*. T. 2. Ed. Foucher, Paris, 1979, 284 pp.

L'ouvrage a un caractère didactique déclaré et il se réfère aux automatisations discrètes industrielles telles que les dispositifs oléopneumatiques et hydrauliques. Le traitement et le développement des problèmes de commandes combinatoires et séquentielles se font progressivement, réalisant continuellement un parallélisme avec les dispositifs électriques.

Parmi les 13 chapitres du livre, les huit premiers chapitres traitent les applications dans le domaine des automatisations discrètes classiques.

Dans les chapitres suivants on présente une introduction concernant la logique programmée, avec l'utilisation des organigrammes, des réseaux Pétri et du Grafcet (chap. IX, X, XI). Le dernier chapitre aborde les automates programmables en tant que solution moderne pour les automatisations industrielles discrètes.

L'ouvrage est utile aussi bien pour les étudiants, pour qui ce livre est un guide permettant la résolution des problèmes d'automatisations industrielles discrètes, que pour les spécialistes surtout à cause du domaine large des problèmes présentés.

C. Huțanu

JARASS L., HOFFMANN L., JARASS A., OBERMAIR G., *Windenergie. Eine systematische Bewertung des technischen und wirtschaftlichen Potentials für die Stromerzeugung der Bundesrepublik Deutschland*. Durchgeführt im Auftrag der Internationalen Energieagentur (I.E.A.) Springer - Verlag, Berlin, Heidelberg, New York, 1981, 272 S.

Das hier besprochene Buch stellt ein auf die Tatsachen der Bundesrepublik Deutschland (B.R.D.) basiertes Studium der gegenwärtigen Benützungsproblemen der Windenergie, dar. Die wichtigsten Kapitel der Arbeit sind die Windenergie innerhalb der zukünftigen Energieversorgung der B. R. D., Determinanten der Windenergiebenützung, die Windverhältnisse in der B.R.D., Umwandlung von kinetischer Strömungsenergie in elektrische Energie, herkömmliches Energieversorgungssystem, SWING, ein Simulationsmodell zur Integration von Windenergie in die Stromversorgung, Brennstoffeinsparung durch Windkraftwerke, Einsparung von Kraftwerkskapazität durch Windkraftwerke.

Die zwei Zusammenfassungen sind zentraler Ergebnisse der Berechnungen des Kapazitätseffekts und der Brennstoffeinsparung, die für den Küstenverbundsystem der B.R.D. durchgeführt werden, gewidmet.

Im Buch wird also der Platz der Windenergie innerhalb der zukünftigen Energieversorgung der B. R. D., zugleich mit den zu beachtenden technologischen und wirtschaftlichen Aufgaben, angedeutet.

Die technischen Probleme der Windenergieproduktion werden für die GROWIAN - Technologie, eine Referenzwindkraftanlage, durchstudiert. Es ist die Rede von den Verhältnissen zwischen Energieerzeugung und die Jahresdurchschnitts-Windgeschwindigkeit, die zeitliche und die räumliche Verteilung der Leistungen, die optimale Aufstellung von Windkraftwerken,

vom Verhältniss zwischen Schnellanlaufzahl und Leistungsbeiwert, vom Zusammenhang zwischen Rotordrehzahl und der Jahresproduktion u.s.w.

Um eine realistische Abschätzung der Auswirkungen der Integration von Windenergie im Energieversorgungssystem zu erzielen wird das umfangreiche Simulationsmodell SWING, welches die jährliche und tägliche Bilanzen sowie die langjährige Entwicklung berechnet, ausgearbeitet.

Obwohl im Studium nur die Winddaten der B. R. D. dargestellt und genützt werden, kann das Datenmaterial und dessen Auswertung als Modell für andere Länder und Gegenden, dienen.

Der extremen Windgeschwindigkeiten und deren Auswirkung auf den Windanlagen wird besondere Aufmerksamkeit, geschenkt.

Dieses wertvolle Buch kann viel zur wirtschaftlichen Benützung der Windenergie im unserem und in anderen Ländern, beitragen.

I. Bergman

GRAY P. R., MEYER R. G., *Analysis and Design of Analog Integrated Circuits*. John Wiley & Sons, Inc., 1977, 683. pp.

The book offers a modern and comprehensive treatment of the analysis and design of analog integrated circuits.

The first five chapters analyse the phenomena that lie at the basis of the functioning of elements and stages in the ICs, the modern technologies of achievement and design considerations.

Chapters 6...9 treat of the structure and behaviour of ICs: the operational amplifiers analysis, feedback and frequency response. They comprise considerations on the design of OAs and voltage regulators.

Chapters 10 deals with the structure and functioning of nonlinear analog ICs: analog multipliers and phase-locked-loop (PLL).

The last chapter (11) is consecrated to noise in the linear ICs.

The book satisfies the highest exigencies, the authors having extensive experience in both industry and teaching.

Particularly useful are the numerical examples which accompany each paragraph and large number of problems to be found at the end of each chapter.

The book proves particularly useful to the students from the electronics faculties, to the specialists dealing in linear integrated circuits and to those who use these circuits.

Vi. Cehan

GIACOLETTO L. J., (red.), *Electronics Designers' Handbook*, Mc. Graw Hill Inc., 1977, 2344 pp.

The book, a second edition completely revised, was written by 28 specialists, professor Giacoletto having the main contribution. The book is divided into 28 chapters: Units, symbols, definitions, formulas and conversion factors, Electrical and electronic properties of materials, Network components, Signal analysis, Circuit analysis, Filters, Attenuators and equalizers, Transmission lines, Vacuum tubes, Solid-state devices, Integrated electronic devices, Power supplies, Small-signal amplifiers, Large-signal amplifiers, Power amplifiers, Signal sources, Servosystems, Wave-shaping circuits, Wave-generating circuits, Digital logic and circuits, Computer-aided design and analysis, Modulation, Receivers, Superheterodyne techniques, Transmitters, Radar and navigational systems, Information theory, Reliability/availability of systems.

The main feature of the book consists in the fact that an enormous area of subjects has been covered in an unitary, clear, elegant and rigorous manner. The electronic engineer will find here not only sistematized informations about the subjects enumerated above but also a carefully selected bibliography at the end of each chapter. Another quality of the book consists in the great number of examples that permit an easier understanding of the different subjects. In fact, the book is useful not only as an aid in electronic design but also as a guide in different fields of electronics, reading of a chapter being equivalent to learning and understanding the essence of the field in the shortest time. To have an idea about the

quantity of information contained in the book, we must show that it has 2,344 pages written with small characters and 1,686 illustrations.

The *Electronics Designers' Handbook* represents a precious help for all those that learn or work in the wonderful field of electronics.

L. Goraș

SANDERSON C. P., *Introduction to Microcomputer Programming*. Newnes-Butterworth, London, 1980, 137 pp.

This book is a simple introduction to programming for users of microcomputers, whether commercial users, teachers or engineers. No previous knowledge of programming, mathematics or electronics is necessary.

In the first chapter there are presented some fundamental notions: program, flowchart, loop, array, subroutine. Also there are explained the main steps we must follow when we write a program.

In the second chapter there are briefly presented the most important programming languages and is explained the meaning of compiler and interpreter.

Chapters 3...7 describe the BASIC language showing some variations which appear at the different suppliers of soft.

Chapters 8 and 9 refers to assembly language and machine-code. There are shown some characteristics of these languages about the arithmetic of negative numbers, the multi-precision arithmetic, the use of BCD code, about the possibilities of using loops, subroutines, and interrupts. The examples are confined to the four most used microprocessors: 6 800, 8 080, z-80, 6 502.

In the last chapter there are some indications about how to verify and test the programs before running them.

This book is helpful for those who work with microcomputers and microprocessors and wish to realize the first programs for these powerful tools.

C. Huțanu

GARLAND H., *Introduction to Microprocessor System Design*. Mc Graw Hill, 1979, 192. pp.

The author presents in the book the essence of microprocessor as well as the principles on which it works and how it can be used. Many specific design examples are used to illustrate the main concepts. Both hardware and software are discussed in the book but assuming some previous background of these matters for the reader.

After a short introduction to the microprocessor technology and evolution, the 8 008, 8 080, z-80, 8 748, 8 086 and z-8000 are presented. The basic microprocessor hardware is next given, followed by methods of expanding the microprocessor systems. Languages concerning microprocessors are presented in two distinct chapters. One of them deals with machine-languages and the other one with assembly and high-level languages. The use of the microprocessor to carry out arithmetic operations is described in a next chapter. In the last two chapters the author describe various analog interfaces as D/A conversion, software timing conversion, software successive approximation conversion, hardware successive approximation conversion and also interface standards.

As the author's idea was, the book must be used in conjunction with the detailed manufacturers' literature available for different specific microprocessors.

L. Goraș

MADDOCK R. I., *Intermediate Electronics*. Vol. 2, Butterworths, London, 1970, 307 pp.

In this volume, the author analyses a wide range of electronic circuits, mainly switching circuits.

The waveforms occurring in such circuits are square waves, sawtooth waves and trains of pulses and rarely sinusoidal waves.

The first chapter presents the transient response of simple CR circuits, diode circuits and transistors.

The following two chapters are concerned with the basic properties of the bistable circuit and multivibrator circuits.

Chapter 4 examines the operation of various waveforms generators: sawtooth, staircase waveforms, very short pulses as applications of the multivibrator family.

Further four chapters cover the principles of logic gates, operational amplifiers and applications to analog computing, d.c. power supplies for transistor equipment and controlled rectification.

Any of this chapter contain many numerical exercises with fully worked solutions and problems with answers, for a good comprehension a theoretical knowledge.

This book is intended to bridge the gap between elementary textbooks and the more academic book available. This text is valuable for students and for technicians.

C. Mărcuță

TURNER C.R., *Real-Time Programming with Microcomputers*. D. C. Heath a Co., Lexington 1978, 160 pp.

This book's area of concern is a particular segment of the total applications activity that is possible with the microcomputer; real-time monitoring and control of external devices.

The real-time segment of a microprocessor-based system is that portion of the total system in which the computer's principal task is to monitor and/or control a time-critical process whose own logical sequence is asynchronous to that of the computer itself.

A simple enumeration of the chapters informs us on the contents of this book; The computer in Real-Time Systems; Servicing an External Device; Interfacing with an External Device; Representing Data; Locating Data; Organizing Data in Memory; Manipulating Data (Transfers, Logical Operations, Arithmetic Operations); Control of Program Flow (Flags and Branching); Implementing Control Structures in the 8080 Assembler; Some Microprocessor Applications.

The author proposes after each chapter exercises to be solved. The work is very useful to the real-time programming with microprocessors system's experts and designers.

We appreciate the usage by the author of the structured programming in the microprocessor's automaton design.

Without disregarding the generality, the author particularises different applications on systems with the microprocessor INTEL 8080.

Al. Valachi

Handbook on Semiconductors. Vol. I, *Band Theory and Transport Properties*, North Holland Publ. Comp.

The aim of this handbook is to provide, in a set of four volumes, a comprehensive coverage of the whole field of semiconductor knowledge. The other three volumes will contain: optical properties of solids (vol. II), materials, properties and preparation (vol. III), device physics (vol. IV). Intended to be a reference source for all workers in the field of semiconductor physics or device development each volume has a separate editor, who is an internationally recognised leader in the field, the entire handbook being coordinated by the well known T. S. Moss, Deputy Director at the famous Royal Signal Radar Establishment, U.K.

The first volume deals, in 17 chapters, with the basic, fundamental properties of crystalline semiconductors: crystal structure, energy bands, transport properties. The last subject has a very large treatment (almost half of the volume), due to its great contribution in establishing the semiconductor properties.

Each chapter is written by a well-known specialist of the domain so that it contains also the latest results of the research. Furthermore, every chapter is provided with a very great number of fundamental and up-to-date references.

The great amount of a well-organised information, the presentation of all points of view in controversial problems, the care for a clear and concise treatment of the subject make the volume to be very useful to all researchers in the field, physicians or engineers,

to all producers and exploiters of semiconductor devices. In the same time the treatment being sufficiently basic, it is very useful, too, for students, both in physics and polytechnical institutes.

I. Bogdan

HAMILTON T.D.S., *Handbook of Linear Integrated Electronics for Research*. McGraw-Hill Book Co. Ltd.

This work is intended for people in a wide range of fields who use electronics in their experimental work and deals concisely with the relevant topics from basic theory to actual integrated circuits and the important principles and factors in practical application. The aim is contained within what are broadly termed linear circuits applications. It omits the large field of communication and digital applications, but this drawback is diminished by the vast and very up-to-date bibliography prepared directly from a computer print out.

After the presentation of a summary of useful basic theory in the first chapter, the author makes some consideration of integrated circuits in the second chapter insisting on the most important of them — the operational amplifier — which reached a very advanced level, with a multitude of models to suit practically every need. The third chapter deals with the most important problem posed to operational amplifier users: the negative feedback. This chapter covers the topic in some detail, in a form suitable for practical application. The following chapter covers the range of the more commonly encountered types of amplifiers realised with operational amplifiers, comprising too some discussion of servosystems theory.

Chapter 5 deals with oscillators, both sinusoidal and function generators, including also some other positive feedback applications.

Chapter 6 covers a number of commonly used functions realised with linear integrated circuits: comparators, multipliers, modulators, phase-locked loops and sample-holds.

Chapter 7 is devoted to power supplies. The author discusses all problems raised by their realisation and underlines the facilities brought by integrated circuits.

The author also deals with some useful discrete devices, as field effect transistors and thyristors (ch. 8), with optoelectronics (ch. 9), and with the extraction of signal from noise (ch. 10).

By its great amount of information and the accuracy of subjects presentation this work fully reaches its aim and may be very useful for electronic engineers too.

I. Bogdan

KATZAN H., *Microprogramming Primer*. Mc Graw-Hill Book Company, 1977. 254 pp.

The reader will find in Herry K a t z a n's book, the introductory principles to microprogramming and emulation topics. The author discuss the relationship between computers, programs and the methods of implementation.

In a first chapter the implementation techniques, the concept of microprogramming and its applications are presented. System structures, the processing unit and the main storage unit both with notations and terminology are next described. A whole chapter is devoted to the organization of the microprogrammed computer and another chapter to microprogram operation as logic-unit operations and microprogramming methods.

The principles of emulation are extensively treated throughout the book. Also the translator and the simulator are presented in two different chapters. At the end of all chapters many exercises, questions and vocabulary lists are given.

One can say about the book that the reader will easily understand from it the microprogramming concepts, being able at last to write microprograms of any complexity.

L. Goras

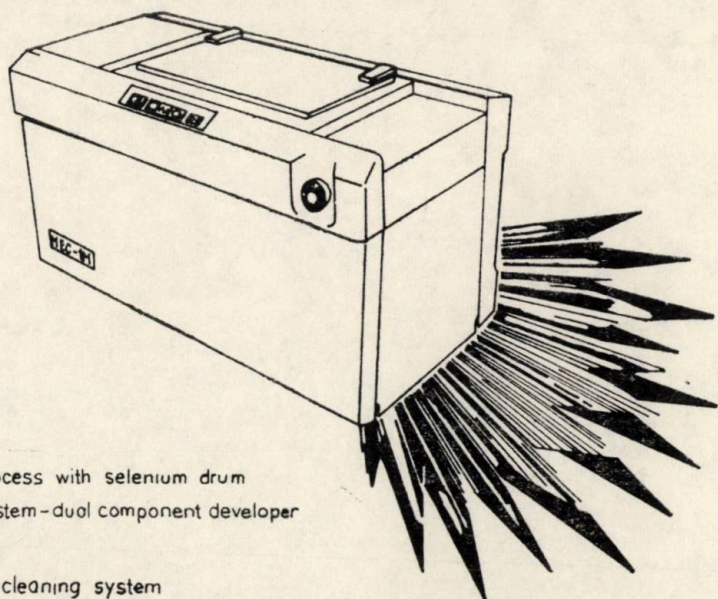


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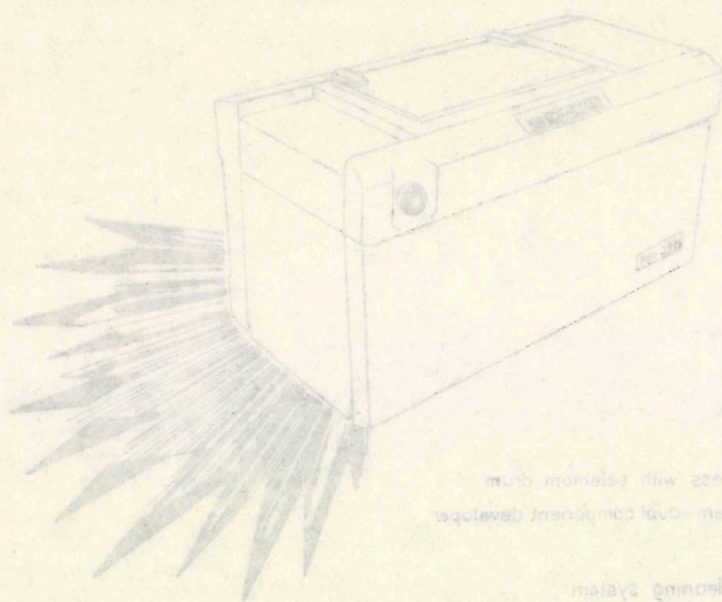
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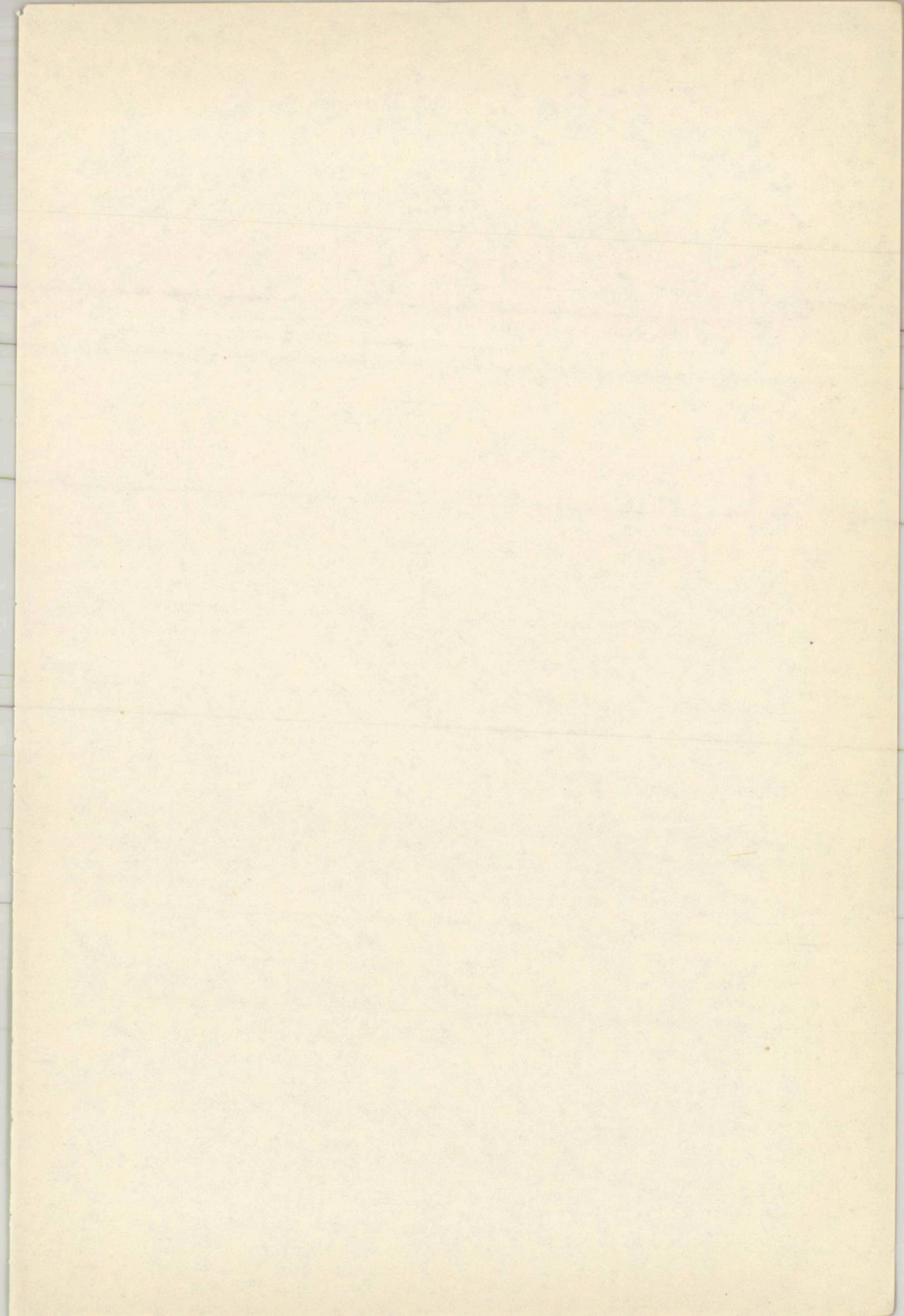
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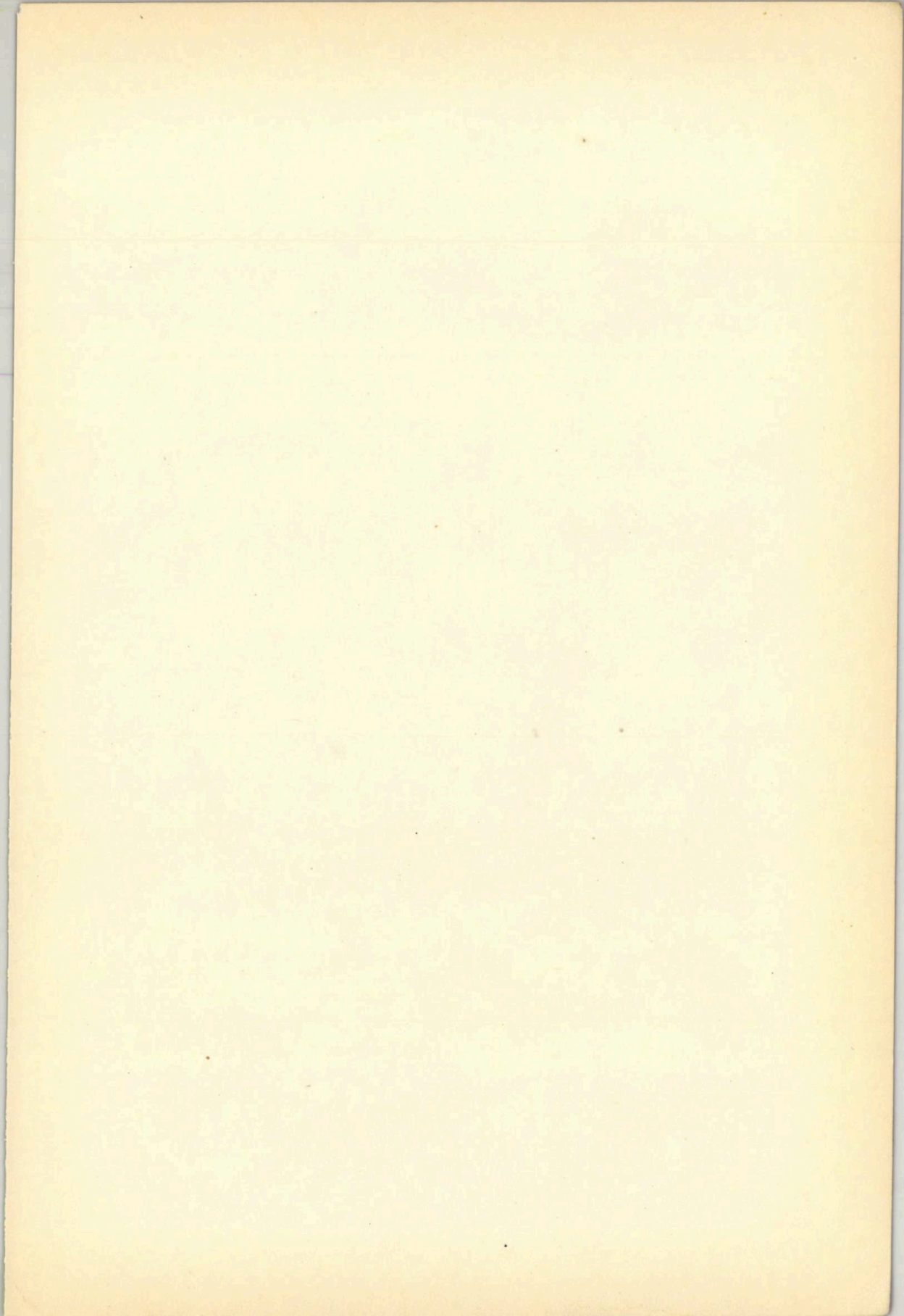
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